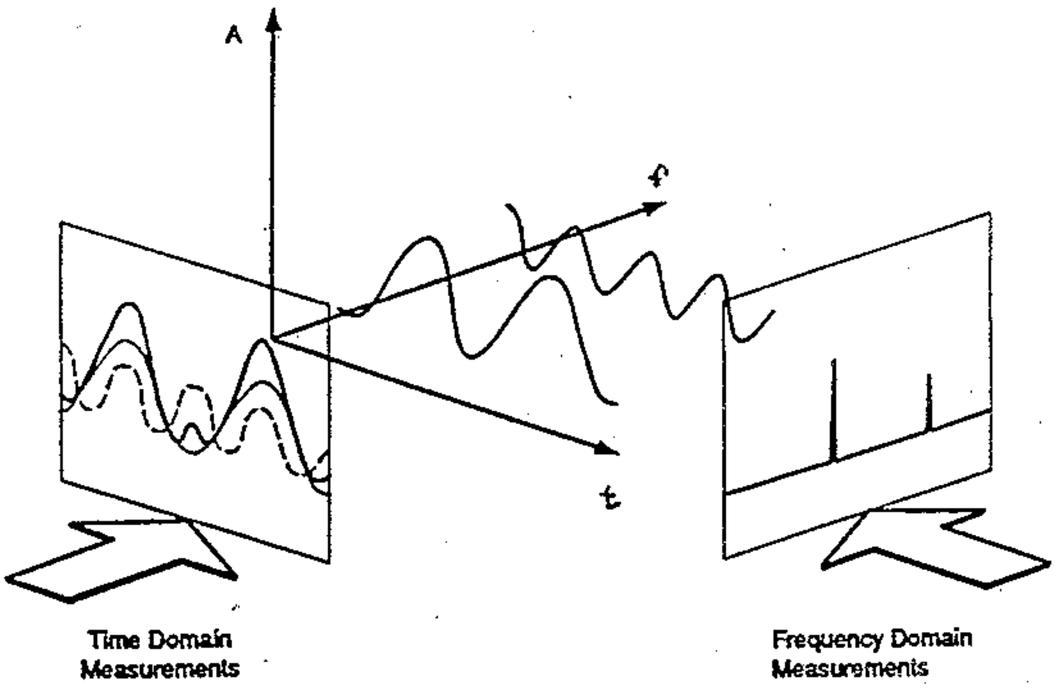
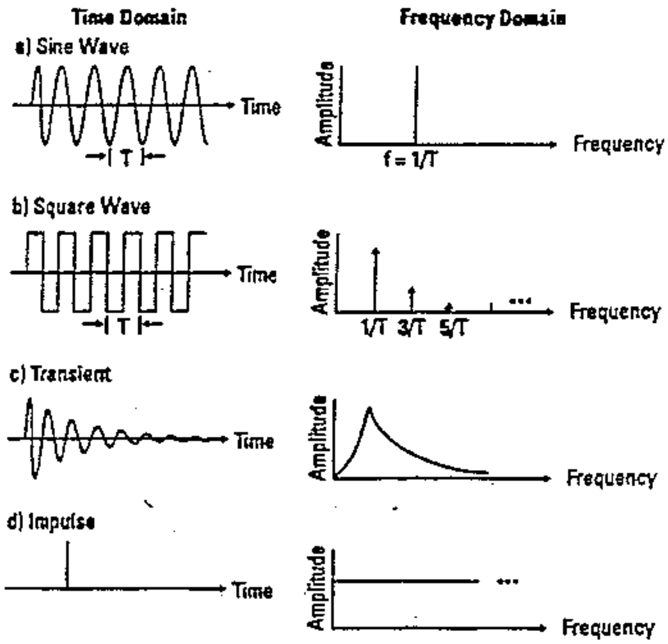
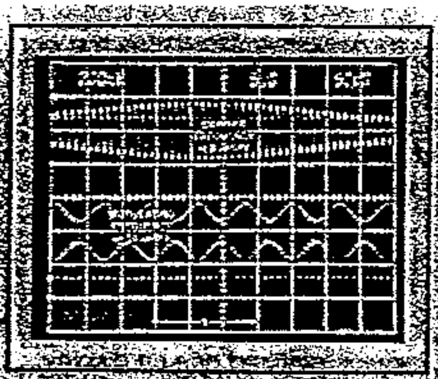
Microwave Measurements in the Time and Frequency Domain



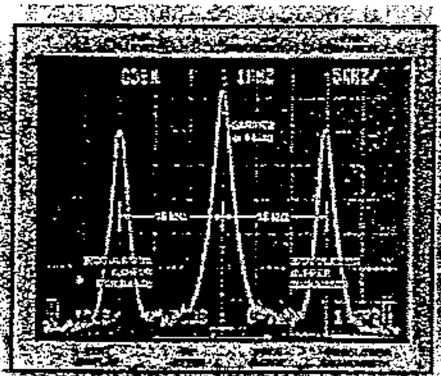
Ralph J. Pasquinelli
Fermilab

Signals Viewed in Both Domains





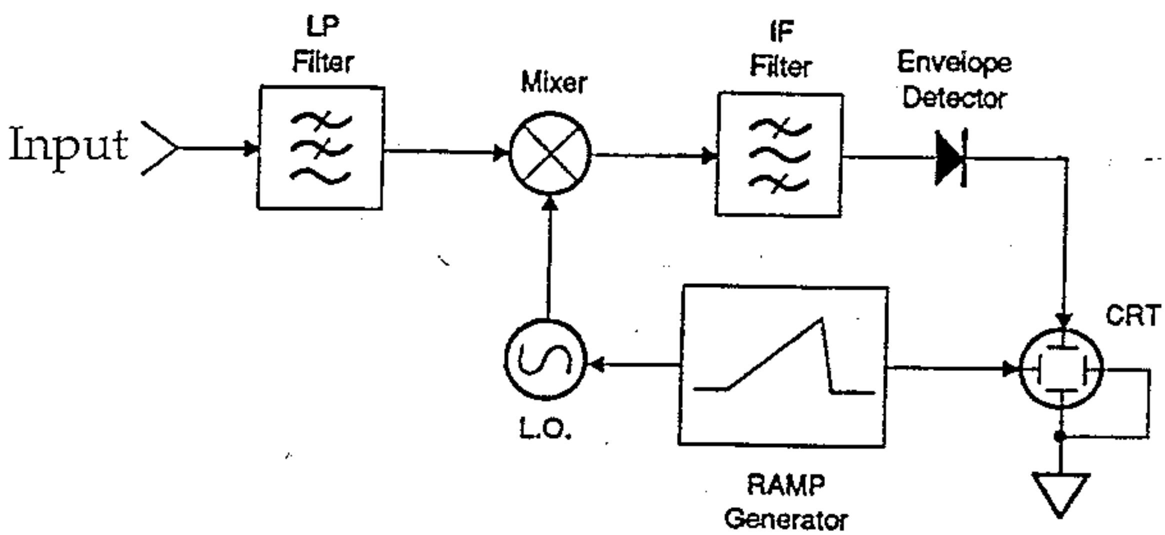
Oscilloscope Waveform: Modulated Carrier at 1 MHz/15 kHz Modulation



Spectrum Analyzer Waveform: Modulated Carrier at 1 MHz, 15 kHz Modulation

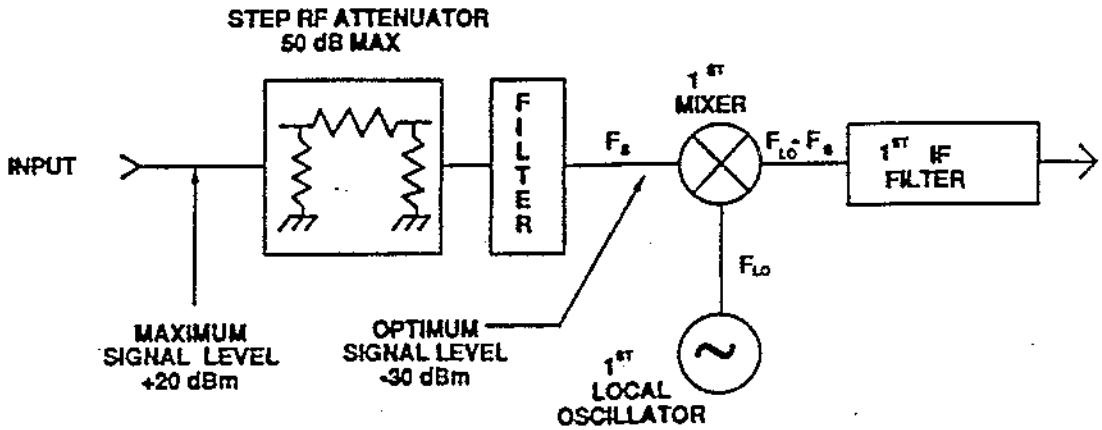
Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer Block Diagram



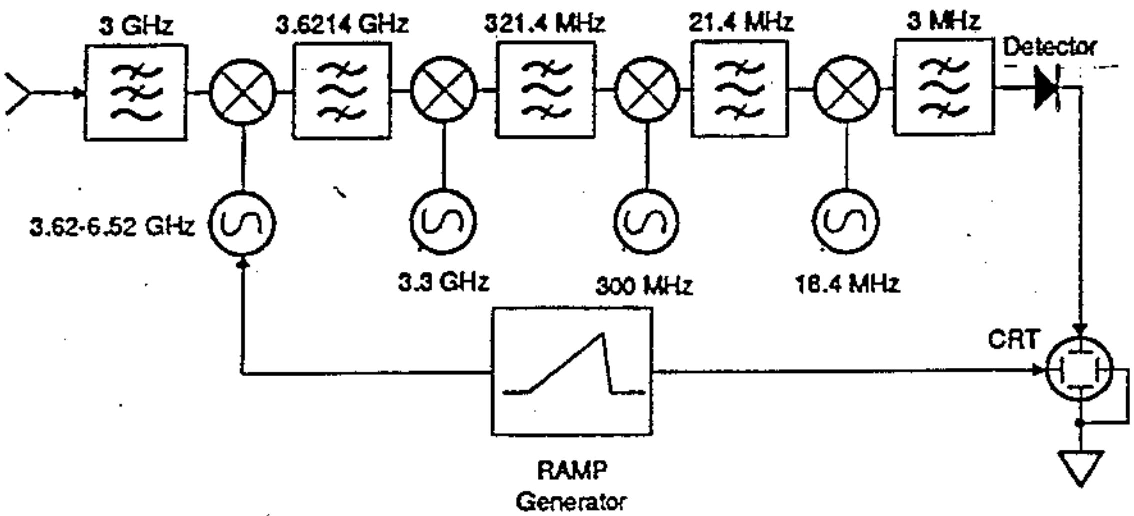
Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer Front End Block Diagram



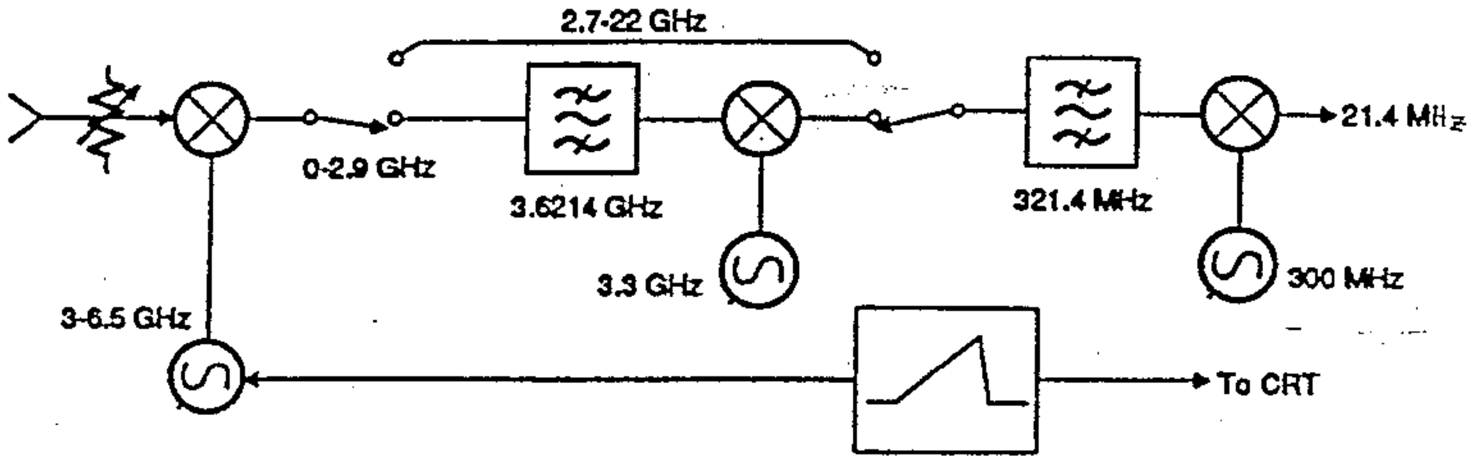
Ralph J. Pasquinelli
Fermilab

Multi-Stage Spectrum Analyzer Block Diagram



Ralph J. Pasquinelli
Fermilab

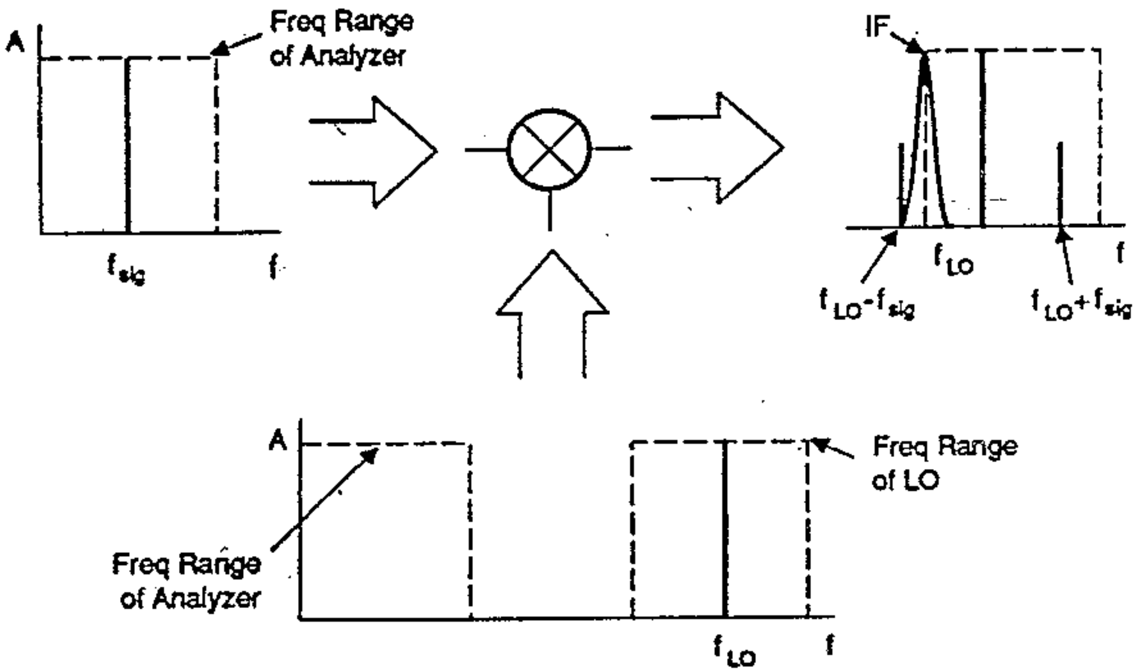
High Frequency Spectrum Analyzer Block Diagram



Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Intermediate Frequency (IF)

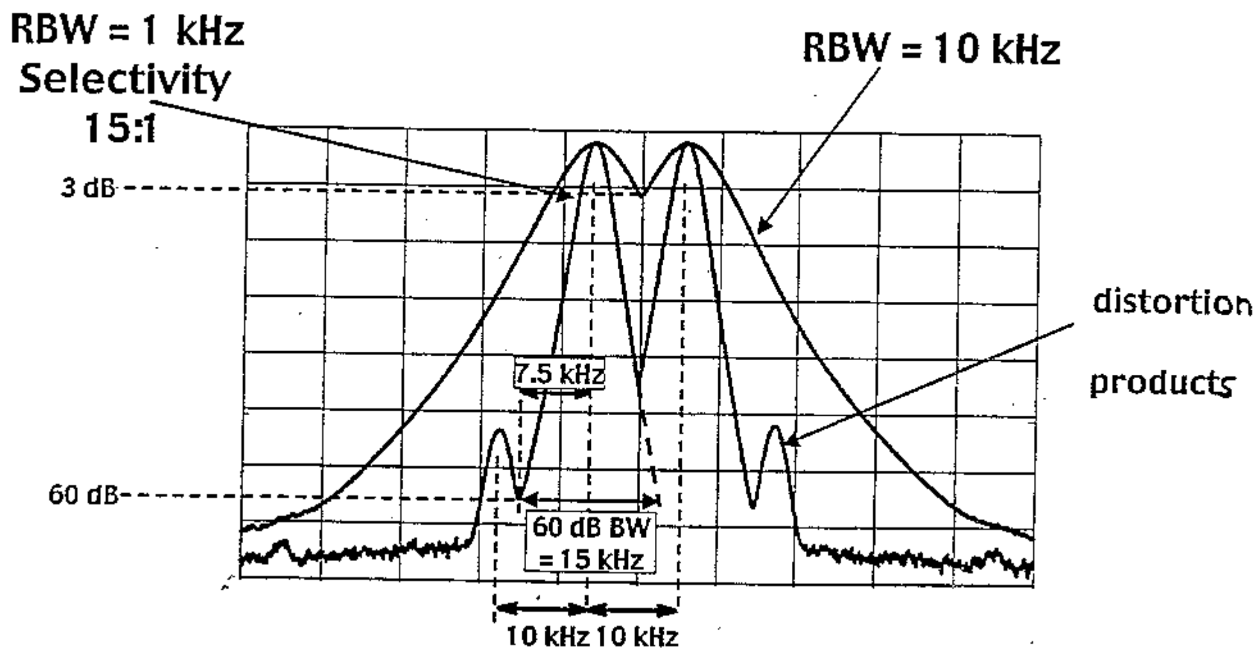
$$f_{IF}=f_{LO}-f_{Signal}$$



Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Resolution Bandwidth (RBW)

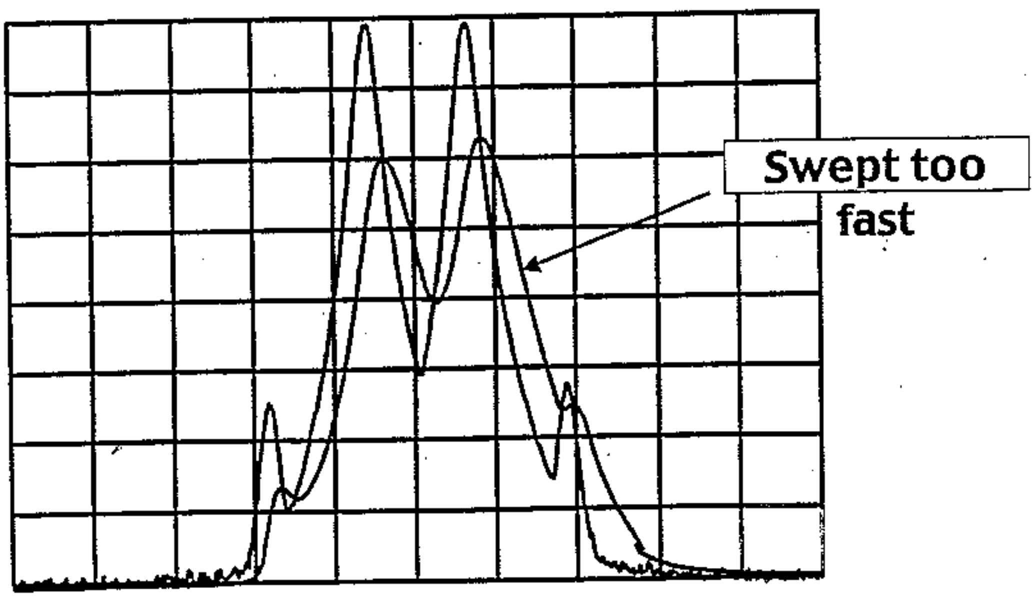
Resolution: RBW Type and Selectivity



Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Sweep Time
and
Calibration Error

Resolution: RBW Determines Measurement Time

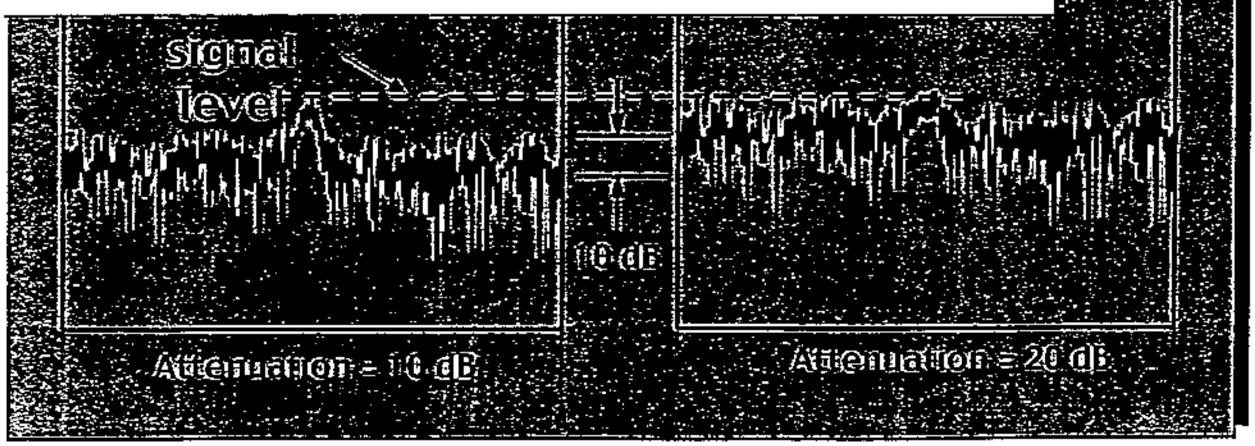


Penalty For Sweeping Too Fast Is An Uncalibrated Display

Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Noise Performance

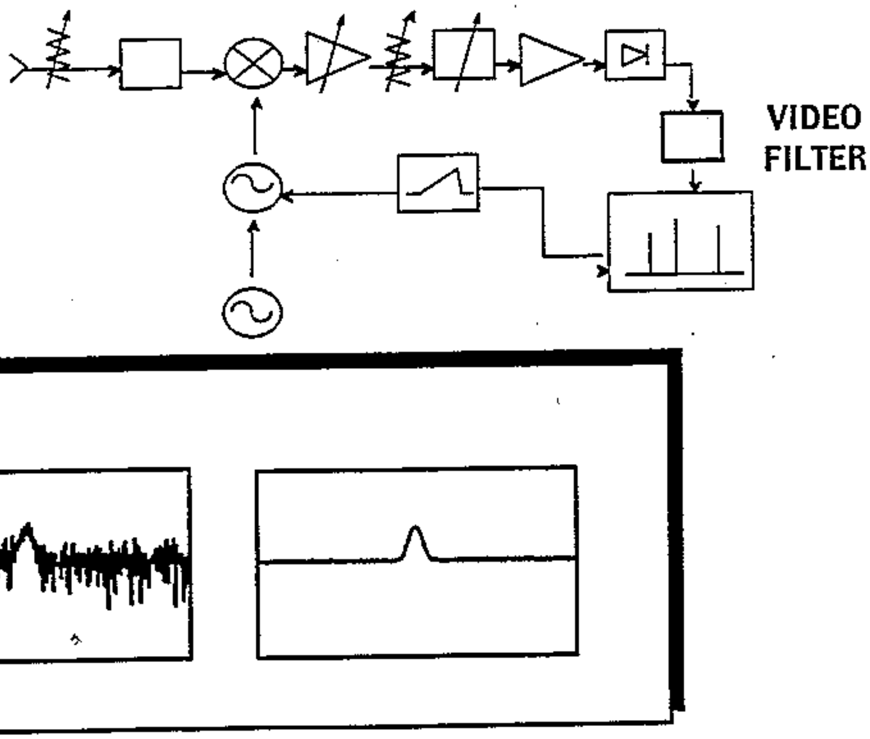
Effective Level of Displayed Noise is a Function of RF Input Attenuation



Signal-To-Noise Ratio Decreases as RF Input Attenuation is Increased.

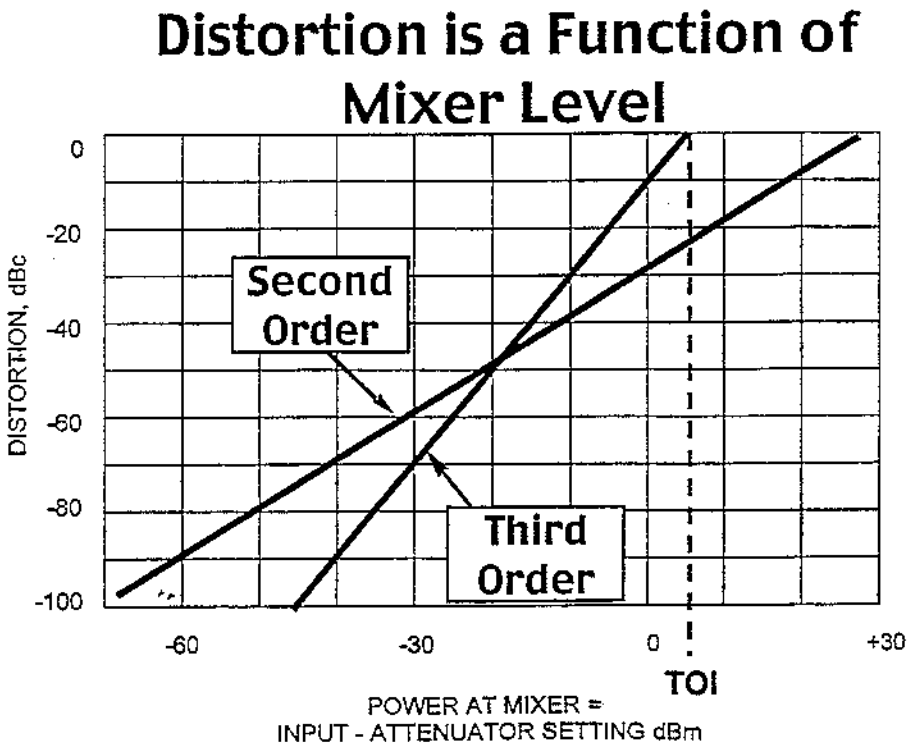
Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Video Bandwidth (VBW)



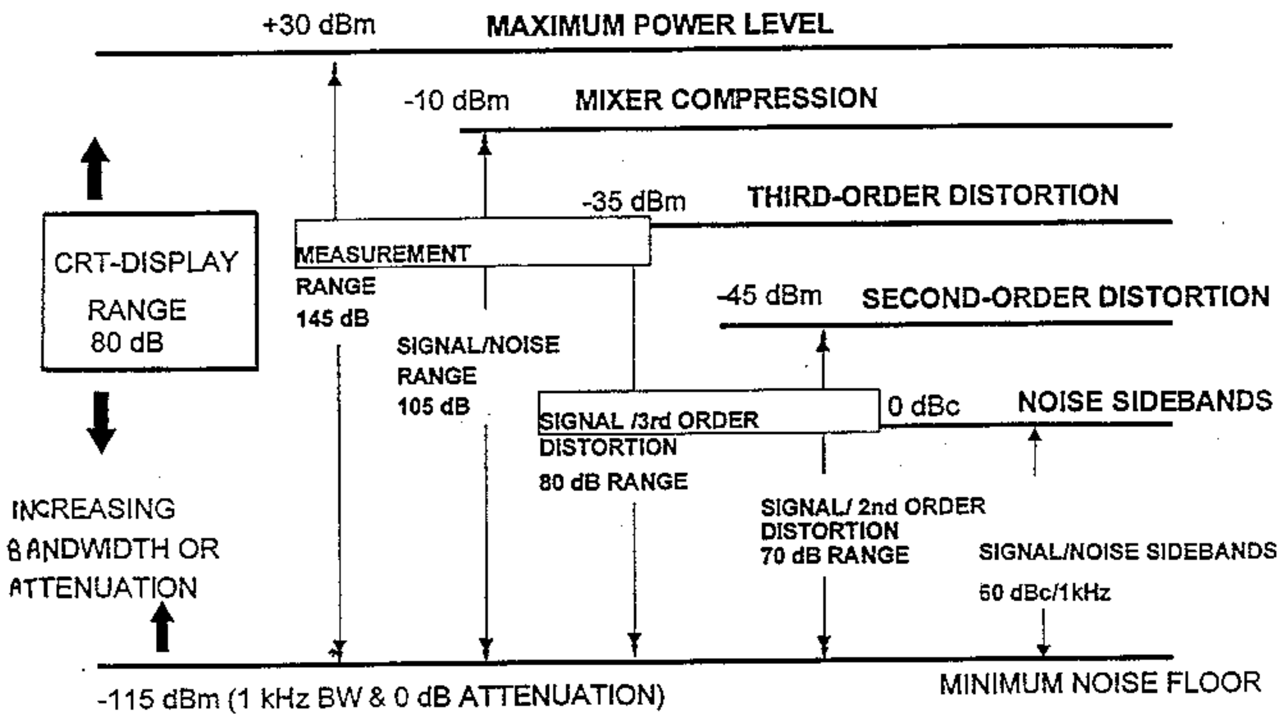
Ralph J. Pasquinelli
Fermilab

Spectrum Analyzer
Distortion and Third Order Intercept



Ralph J. Pasquinelli
Fermilab

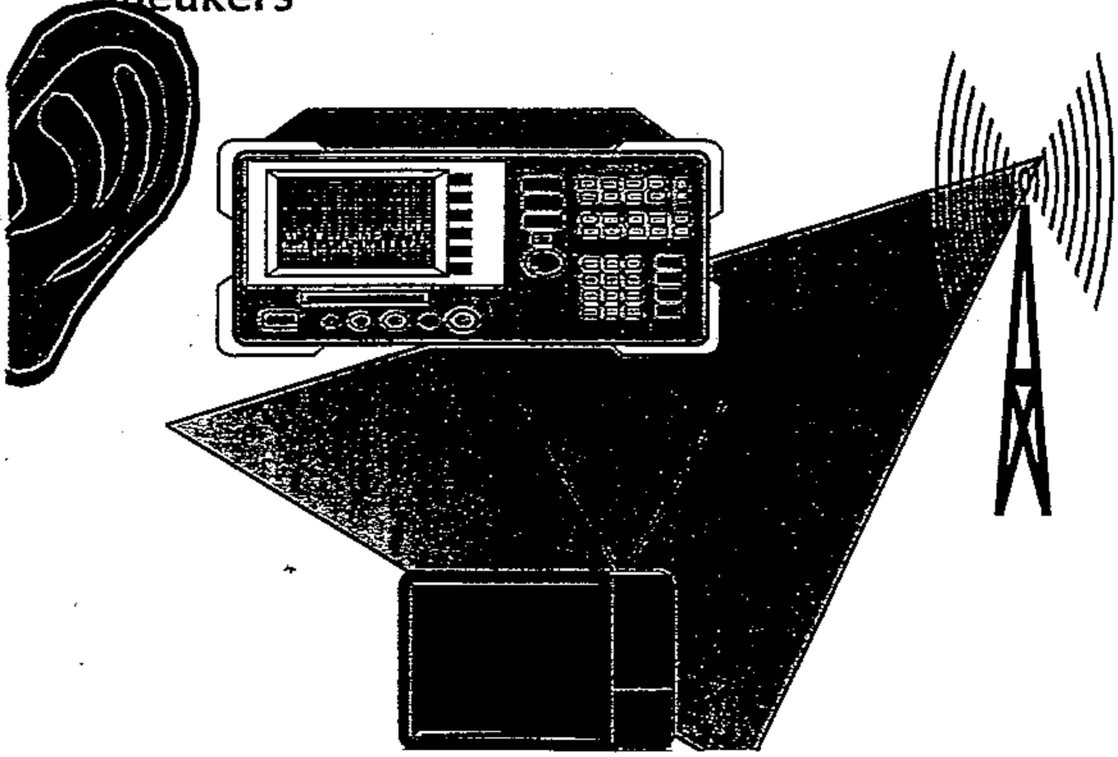
Spectrum Analyzer
Dynamic Range



Ralph J. Pasquinelli
Fermilab

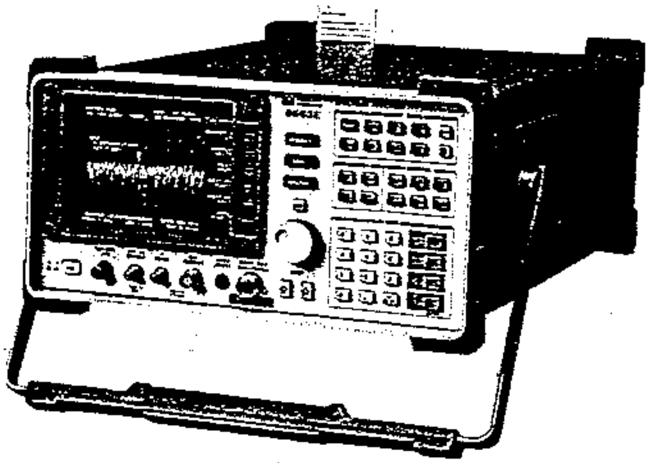
Spectrum Analyzer
Zero Span and Modulation Detectors

Modulation Measurements: AM/FM Detector with Speakers



Ralph J. Pasquinelli
Fermilab

- Continuous 30 Hz to 2.9, 6.5, 13.2, 26.5, 40, or 50 GHz sweeps
- Resolution bandwidths of 1 Hz to 100 Hz digitally implemented for measurement speed
- Low phase noise and wide dynamic range
- Precision timebase and 1 Hz counter resolution
- Adjacent channel power, channel power, carrier power, and gated video measurements standard
- MIL-T-28800 rugged
- Check out the new specifications on HP 8560 E-series



HP 8563E

HP 8560 E-Series Spectrum Analyzers

The HP 8560 E-series portable spectrum analyzers offer the measurement capabilities and performance traditionally found only in larger, more expensive benchtop analyzers. These spectrum analyzers combine outstanding phase noise, sensitivity, 1 Hz resolution bandwidths, and wide dynamic range in a MIL-rugged package built to withstand harsh environmental conditions.

Capabilities for RF Communications

The ability to measure adjacent-channel power (ACP) on today's wireless telephones, pagers, and other transmitters is critical in both R&D and manufacturing. The HP 8560 E-series spectrum analyzers offer a complete solution for ACP testing of burst carrier signals using digital modulation such as is used in NADC-TDMA, GSM, DECT, CT2-CAI, PDC and PHP system. Many of the implementation difficulties of the established standards have been addressed, providing fast, accurate, and easy-to-use ACP measurement capability.

Another standard feature is the ability to measure from .10 to 99.99 percent occupied bandwidth.

Time-gated signal analysis is another standard feature that allows you to easily measure time-varying signals such as pulsed RF, time-division multiple access, interleaved, and burst-modulated. The HP 85902A burst carrier trigger can supply a TTL trigger signal.

HP 8560 E-series specifications have been enhanced. Now, you can get better phase noise, sensitivity, dynamic range and frequency response from this high performance portable spectrum analyzers family.

The new HP 8562E spectrum analyzer provides a 13.2 GHz frequency range with increased dynamic range and third-order intercept (TOI) capability. This allows wireless-communication engineers to test high-performance components in burst operation systems.

With HP 85672A spurious response measurements utility, you can use HP 8560 E-series spectrum analyzer to make fast and easy spurious response test.

For more information on RF communications measurement capabilities, refer to page 477.

Fast Digital Resolution Bandwidths

Digitally implemented resolution bandwidths of 1, 3, 10, 30, and 100 Hz allow the HP 8560 E-series spectrum analyzers to sweep from 3 to 600 times faster than is possible with comparable analog filters. A narrow 5:1 shape factor allows you to view close-in, low-level signals easily. Digital bandwidths also provide the spectrum analyzer with a full 100 dB on-screen calibrated display.

PC Software Utility for HP 8560 Series

With the new Screen Capture PC software utility you can "capture" your measurement results and transfer analyzer screen images or trace data over HP-IB or RS-232 interfaces to a personal computer.

Screen Capture for HP Analyzers can be obtained free from the World Wide Web at http://www.tmo.hp.com/.

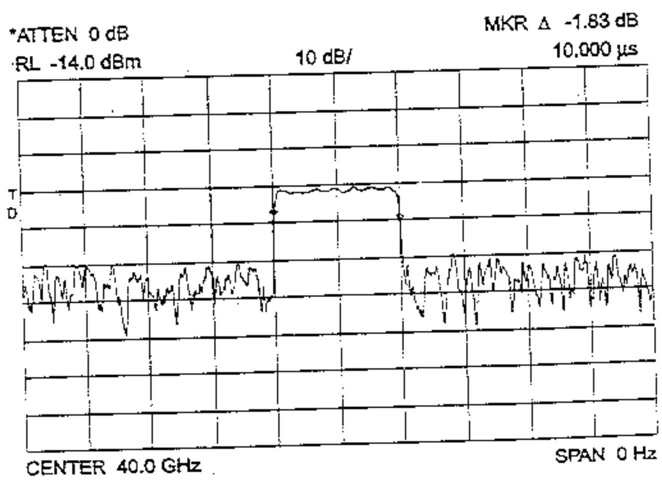
Precision Frequency and Amplitude

Measure frequencies accurately using the built-in frequency counter. A standard precision frequency reference, with an aging rate of 1×10^{-7} per year, and 1 Hz counter resolution provide confidence in measurement accuracy. At 1 GHz, frequency accuracy of ± 135 Hz after a 15-minute warmup is achieved.

Amplitude measurement uncertainty can be reduced using the new amplitude correction (AMPCOR) feature. AMPCOR allows you to enter up to 200 amplitude correction points to compensate for sources of amplitude uncertainty, such as cable losses, preamplifier gain, and spectrum analyzer frequency response. After developing a table of correction data, amplitudes that have been referenced to a power meter can be read directly on the spectrum analyzer display.

Digitized, Fast Time-Domain Sweeps

Add digitization to fast time-domain (zero span) sweeps with Option 007. Use markers, trace math, trace storage, and get hardcopy output, for measurements such as rise/fall times, pulse widths, and time between events.



With Option 007 markers can be used even with the fastest time-domain (zero-span) sweep times.

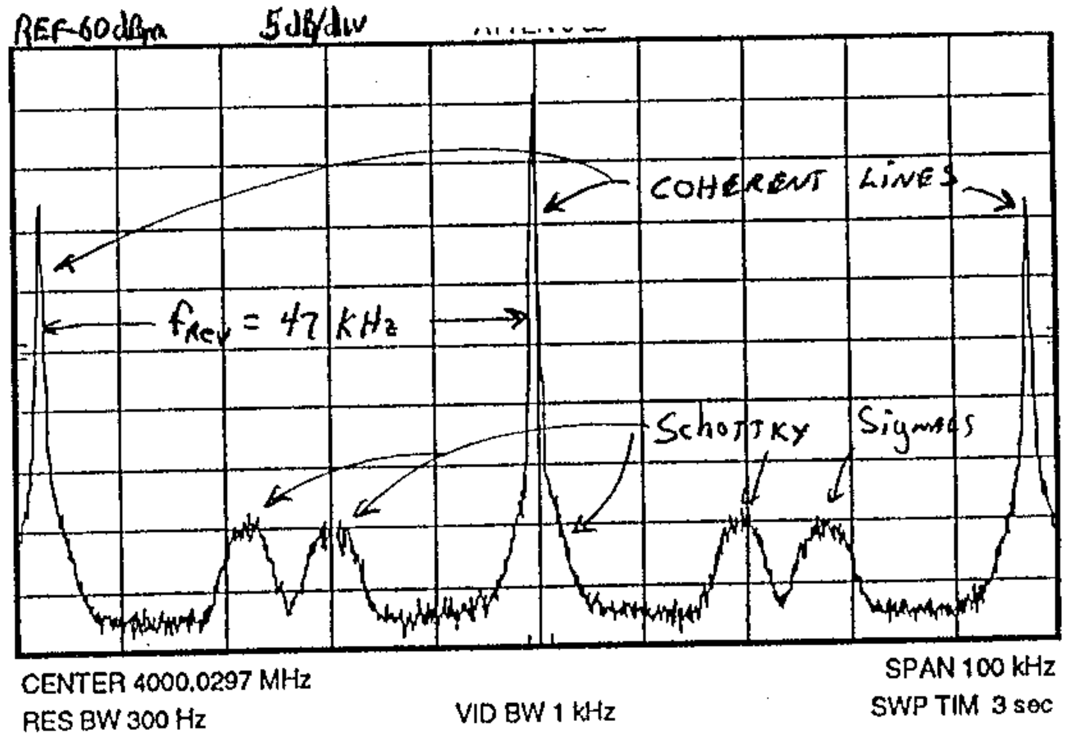
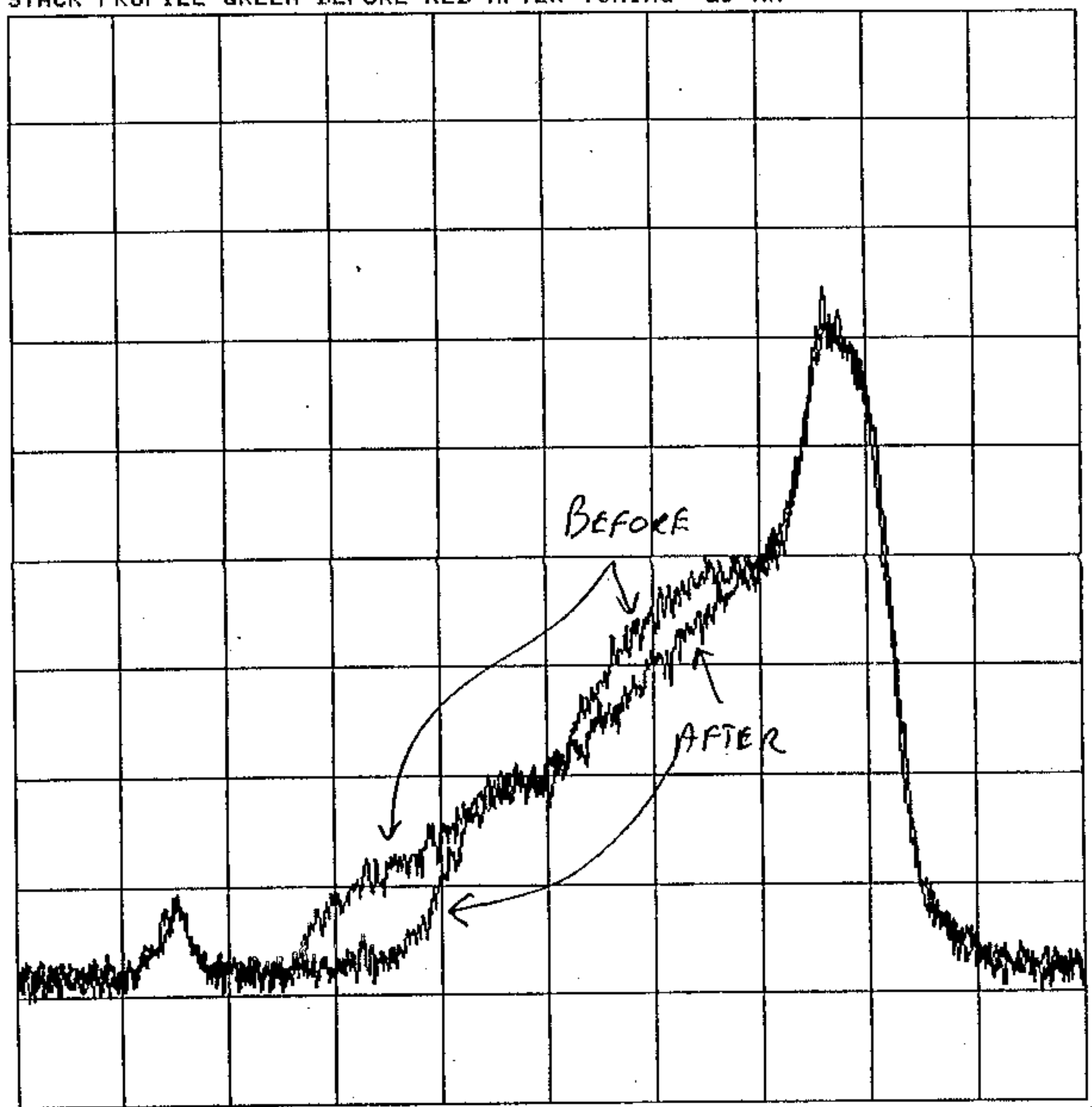


Figure 5. Typical Tevatron vertical bunched beam spectrum at 4 GHz as measured by the vertical proton pickup.

Console Location 3,
Pbar SA Plot

22-MAR-1995 13:21

STACK PROFILE GREEN BEFORE RED AFTER TUNING 29 MA



Start Freq 79.21000001 MHz

Stop Freq 79.26000001 MHz

03/22/95 1306

Scale 10 dB/div

Atten 0 dB

Swp 1 sec

Vid BW 300 Hz

Res BW 300 Hz

Ref Lvl -30 dB

VID AVG

Console Location 1,
Emittance Measurement / Traces

10-DEC-1994 23:46

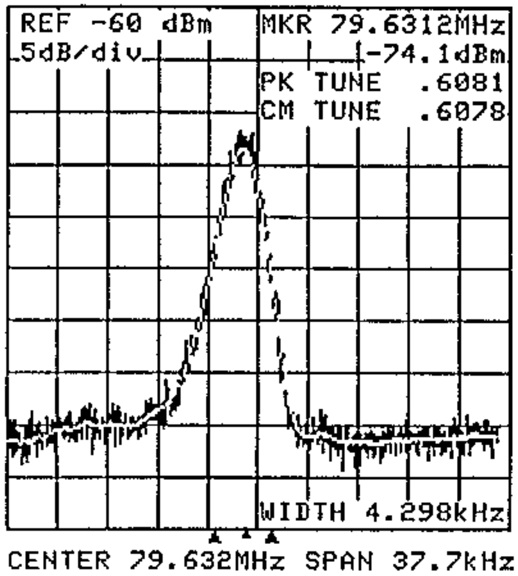
ACCUMULATOR

12/10/94 2343

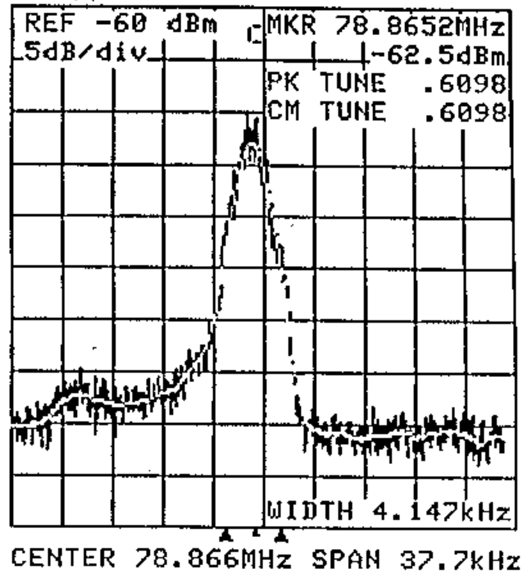
BEAM CURRENT	66.36 mA
STACKING RATE	6.117 mA/h
HARMONIC NUMBER	126
REV FREQUENCY	628958 Hz
DELTAP/P	.1951 %
HORIZONTAL TUNE	.6088
VERTICAL TUNE	.6072
HORZ CHROMATICITY	-.0607
VERT CHROMATICITY	-.1879

Correct Chromaticity
CALCULATION 10/06/92

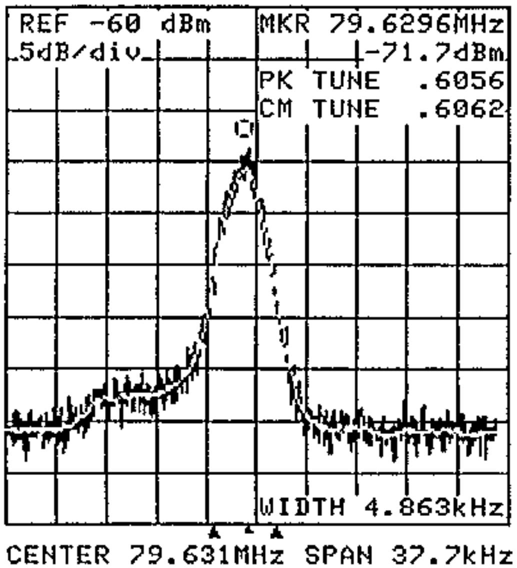
"Lower" HORIZONTAL SIDEBAND



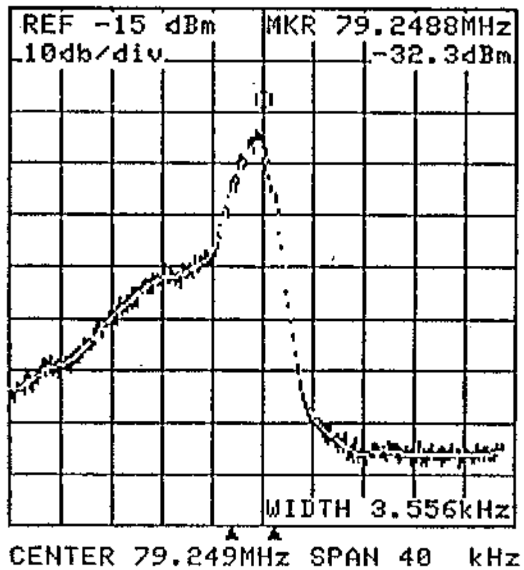
"Upper" HORIZONTAL SIDEBAND



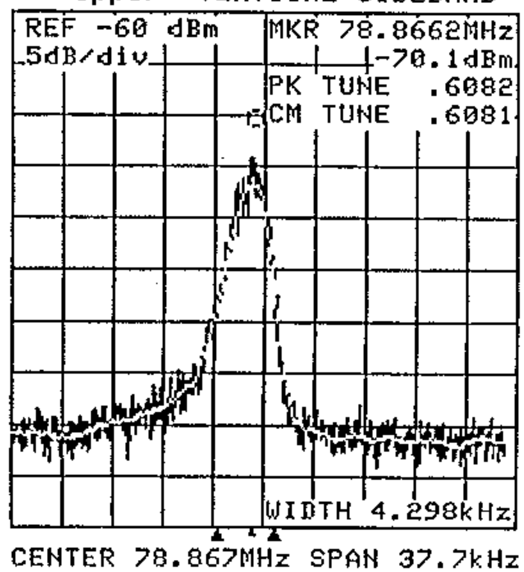
"Lower" VERTICAL SIDEBAND



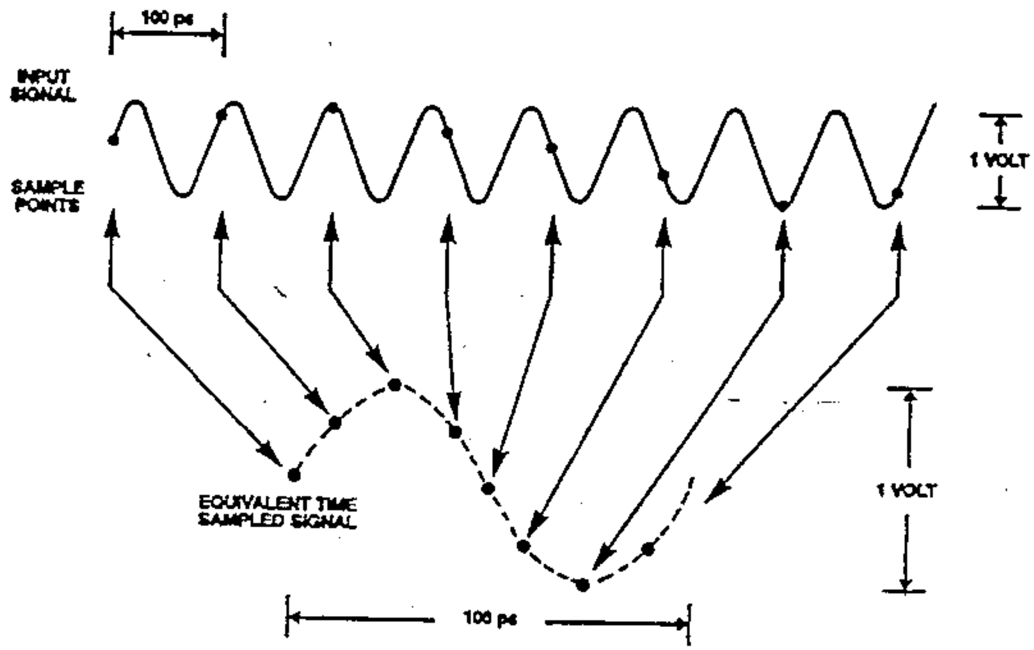
LONGITUDINAL SCHOTTKY



"Upper" VERTICAL SIDEBAND

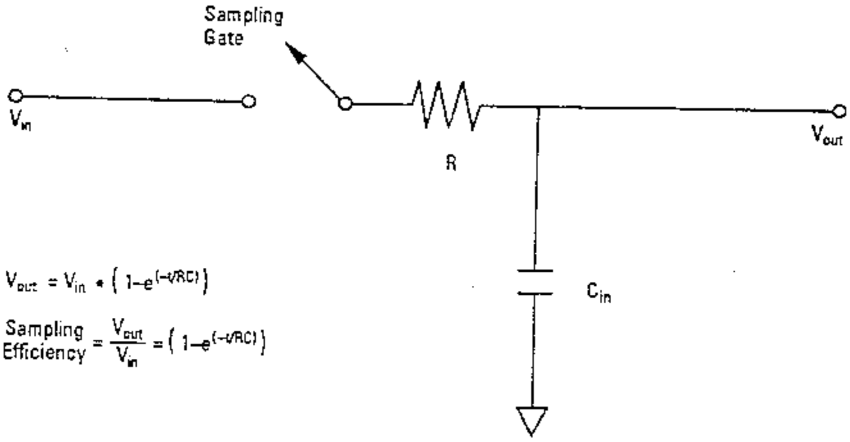


Sampling Oscilloscope



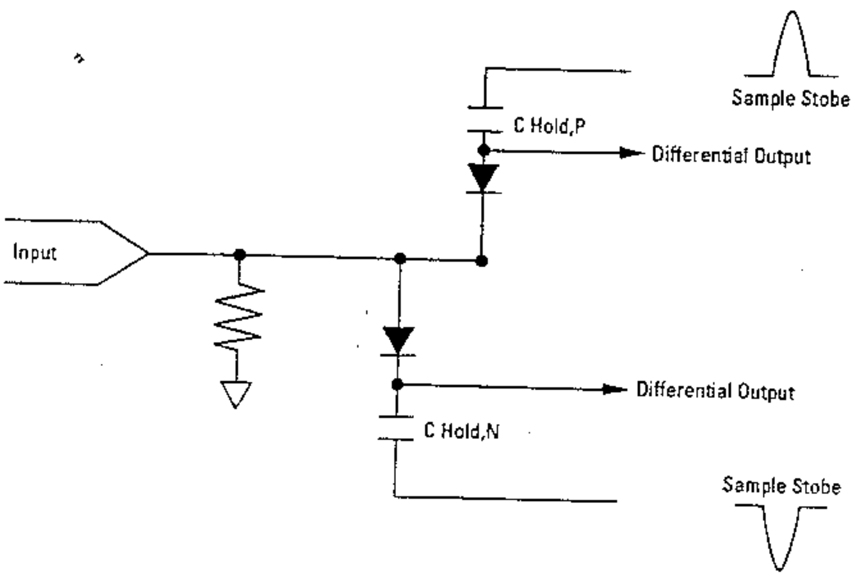
Ralph J. Pasquinelli
Fermilab

Sampling Scope Front End



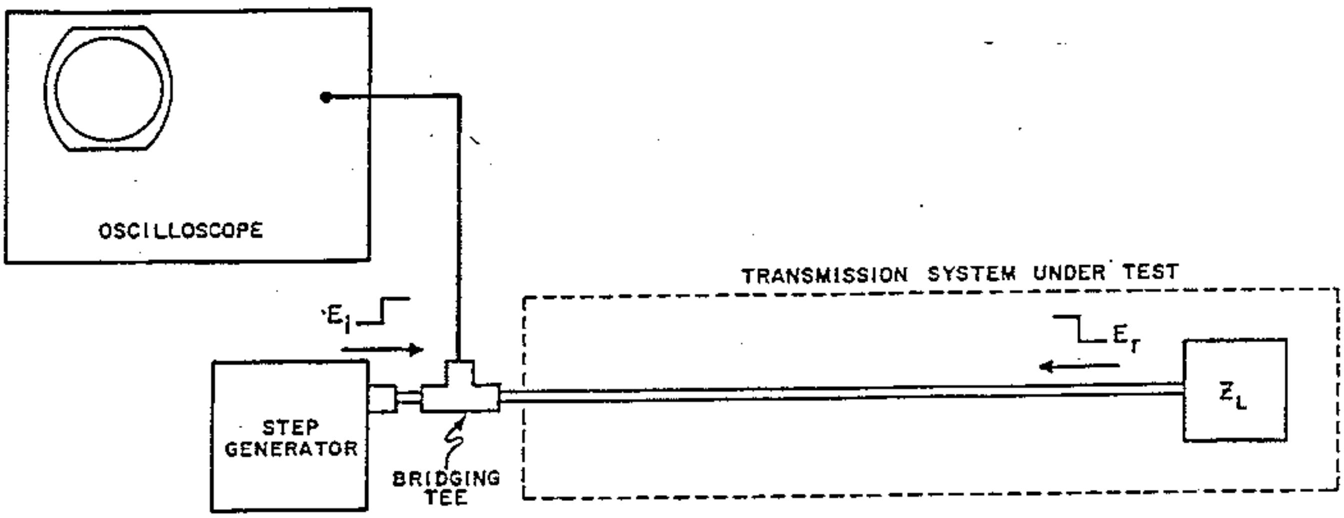
$$V_{out} = V_{in} * \left(1 - e^{(-t/RC)}\right)$$

$$\text{Sampling Efficiency} = \frac{V_{out}}{V_{in}} = \left(1 - e^{(-t/RC)}\right)$$

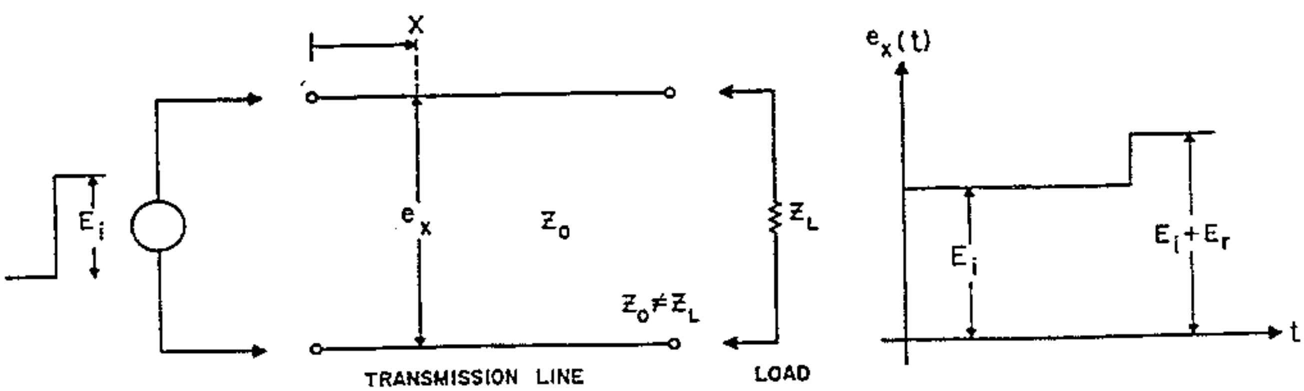


Ralph J. Pasquinelli
Fermilab

Time Domain Reflectometer (TDR)



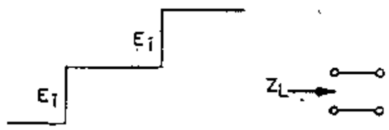
Once the incident and reflected voltages are measured on the oscilloscope, the reflection coefficient and impedance of the mismatch may be calculated.



$$\rho = E_r/E_i = (Z_L - Z_o)/(Z_L + Z_o)$$

Ralph J. Pasquinelli
Fermilab

TDR Displays for Resistive Loads



(A) OPEN CIRCUIT TERMINATION ($Z_L = \infty$)

(A) $E_r = E_i$ THEREFORE $\frac{Z_L - Z_o}{Z_L + Z_o} = +1$

WHICH IS TRUE AS $Z_L \rightarrow \infty$

$\therefore$ Z = OPEN CIRCUIT

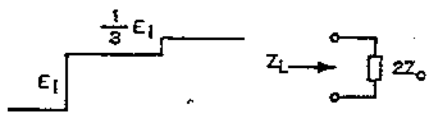
(B) SHORT CIRCUIT TERMINATION($Z_L = 0$)

(B) $E_r = E_i$ THEREFORE $\frac{Z_L - Z_o}{Z_L + Z_o} = -1$

WHICH IS ONLY TRUE (FOR FINITE Z_o)

WHEN $Z_L = 0$

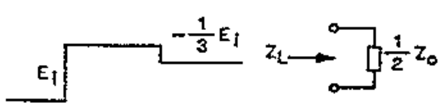
$\therefore$ Z = SHORT CIRCUIT



(C) LINE TERMINATED IN $Z_L = 2Z_o$

(C) $E_r = +\frac{1}{3}E_i$ THEREFORE $\frac{Z_L - Z_o}{Z_L + Z_o} = +\frac{1}{3}$

AND $Z_L = 2Z_o$



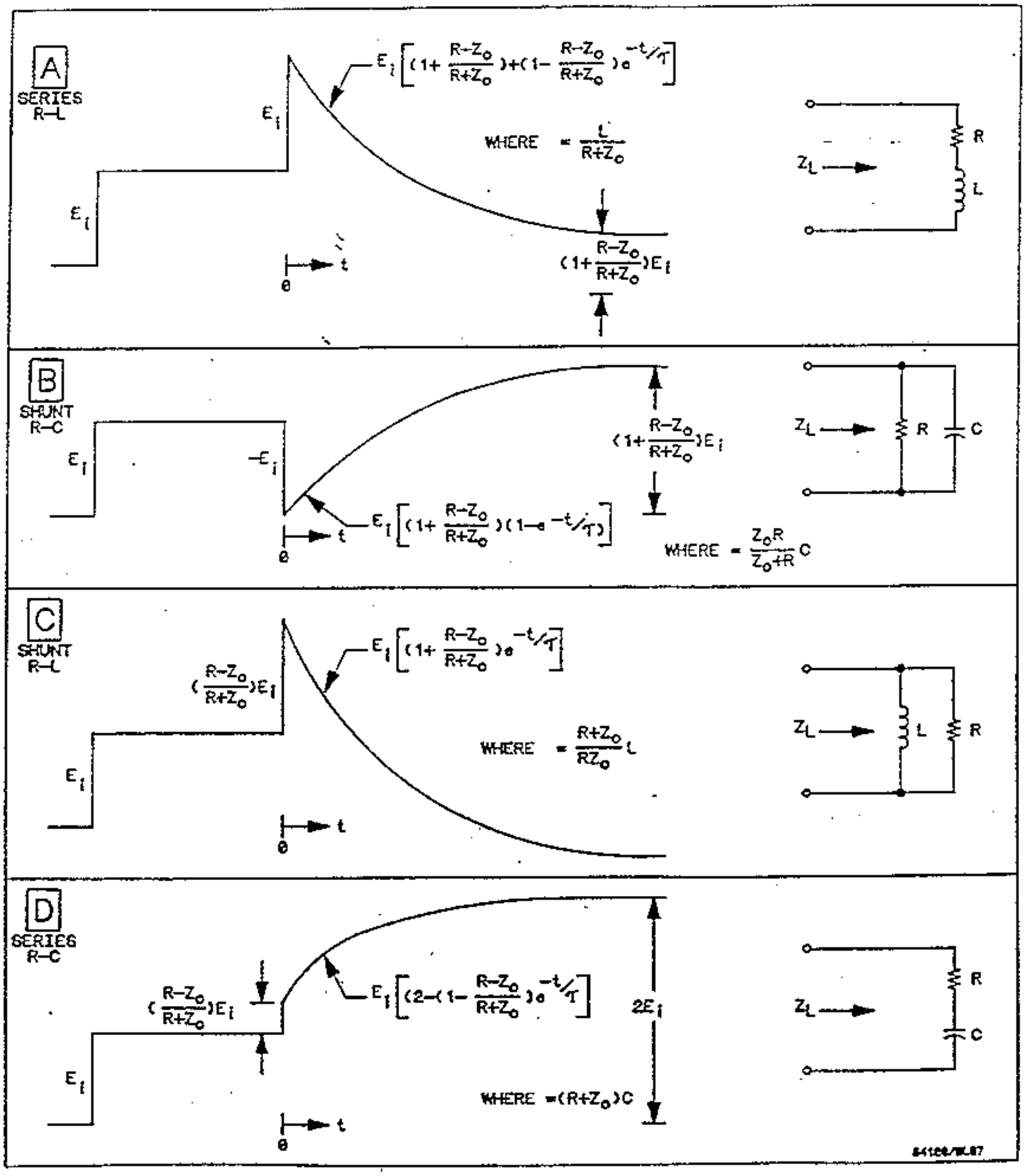
(D) LINE TERMINATED IN $Z_L = \frac{1}{2}Z_o$

(D) $E_r = -\frac{1}{3}E_i$ THEREFORE $\frac{Z_L - Z_o}{Z_L + Z_o} = -\frac{1}{3}$

AND $Z_L = \frac{1}{2}Z_o$

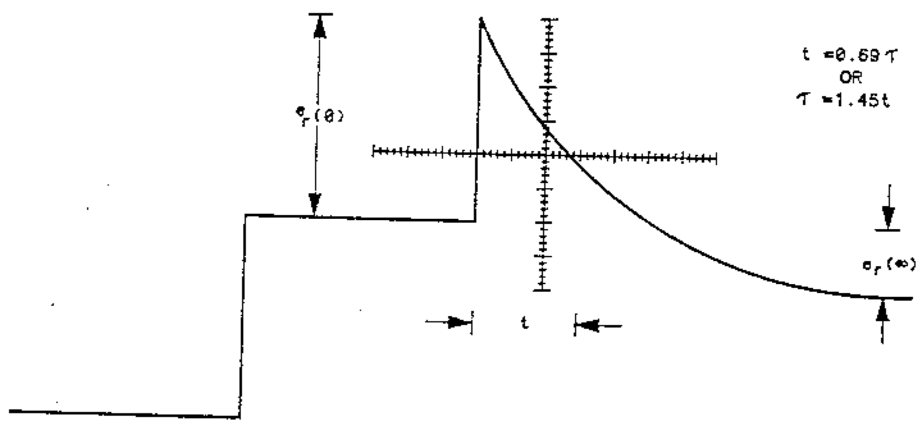
Ralph J. Pasquinelli
Fermilab

TDR Displays for Complex Loads



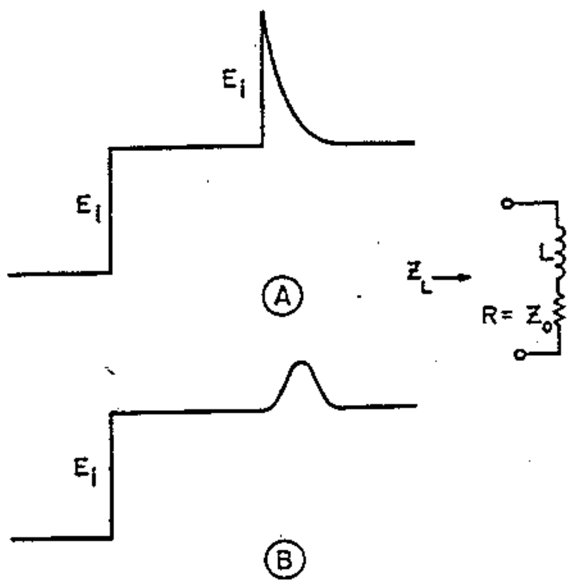
Ralph J. Pasquinelli
Fermilab

TDR Measurement of Time Constants



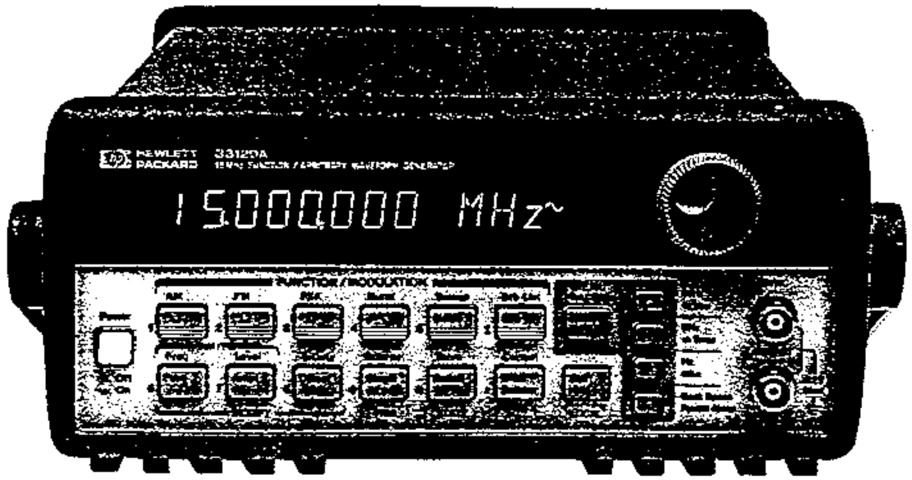
Ralph J. Pasquinelli
Fermilab

TDR Displays with Limited Bandwidth



Ralph J. Pasquinelli
Fermilab

- 15 MHz sine- and square-wave outputs
- Sine, triangle, square, ramp, noise, and more
- 12-bit, 40 MSa/s, 16k deep arbitrary waveforms
- Direct digital synthesis for excellent stability
- Linear and log sweeps built-in
- AM, FM, FSK, and burst modulation built-in
- HP-IB and RS-232 interfaces both standard
- Optional MS Windows Arb software



HP 33120A

HP 33120A Function/Arbitrary Waveform Generator

The HP 33120A is a high-performance, full-function 15 MHz synthesized function generator. It features sine, triangle, square, ramp, and noise waveforms, a 12-bit, 40 MSa/s, 16k-deep arbitrary waveform generator, and both internal sweep and modulation capabilities. The HP 33120A is ideal for both bench and system applications. Both HP-IB and RS-232 interfaces are standard, as is a full three-year warranty. All this for a surprisingly affordable price.

Unprecedented Functionality

The HP 33120A will fill all your basic signal source needs by giving you a full complement of standard functions. But this source goes beyond the basics. You get both linear and log sweeps to 15 MHz, plus full-modulation capabilities. AM, FM, FSK, and burst modulation are just a button push away. You can internally modulate with any of the standard waveforms, including Arb. You can even use an external source for AM, FSK, and burst modulation, if desired. Finally, you get near-infinite custom waveform capability with the inclusion of a 12-bit, 40 MSa/s, 16k-deep arbitrary waveform generator.

Superb Performance

The performance of the HP 33120A was designed in, not left out. This means that you get clean, low-distortion sine waves, fast rise- and fall-time squarewaves, and linear triangle and ramp waveforms. Further, due to the latest direct digital synthesis techniques utilized in the HP 33120A, you can get down to 10 μHz frequency resolution.

Built-In Versatility

You will find that the HP 33120A will fit equally well into your bench or your system applications. Designed with the bench user in mind, operation of the HP 33120A from the front panel is straightforward and intuitive. The inclusion of a knob makes adjusting frequency, amplitude, and offset extremely convenient. Or enter these values directly. You can even enter amplitude values directly in V peak-to-peak, V rms, or dBm. For system applications, the HP 33120A includes both HP-IB and RS-232 interfaces standard, and uses commands that are in total compliance with the Standard Commands for Programmable Instrumentation (SCPI).

Quality and Reliability

Not only does the HP 33120A offer you performance and features unheard of at this price, you also get the advantages of owning Hewlett-Packard. A full three-year warranty is standard with the HP 33120A. The rugged construction and conservative design of the HP 33120A ensures many years of trouble-free operation. Just as price was designed out of the HP 33120A, quality and reliability were designed in.

Option 001 Phase Lock Loop

Option 001 adds a high-stability timebase, the ability to lock to an external timebase, and the ability to phase lock two or more HP 33120A's together. This option is especially useful if your application requires higher-frequency stability and accuracy, if you need to lock to an external-frequency standard, or if you need two or more phase-locked outputs.

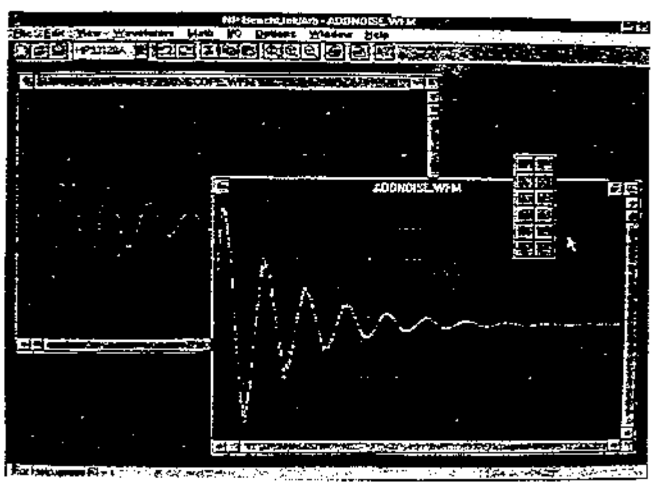
HP BenchLink Arb Software Helps the 33120A Work for You

HP BenchLink Arb lets you use your Windows-based PC to easily create and edit arbitrary waveforms for output on the HP 33120A.

HP BenchLink Arb software application lets you create waveforms in a variety of ways:

- Select and edit a standard waveform from the HP BenchLink Arb library, and change its amplitude and frequency characteristics as desired.
- Use HP BenchLink Arb's drawing tools to draw and edit your own custom waveform.
- Bring in and edit a waveform captured or created elsewhere.

HP BenchLink Arb accepts time/voltage pairs in ASCII format, or you can use waveforms captured with HP BenchLink Scope and an HP oscilloscope. Once your waveform is ready, downloading to the HP 33120A generator is simple. Make your arbitrary waveforms quickly and easily with HP BenchLink Arb.



HP 34811A BenchLink Arb lets you take control of arbitrary waveforms on the HP 33120A function/arbitrary waveform generator.

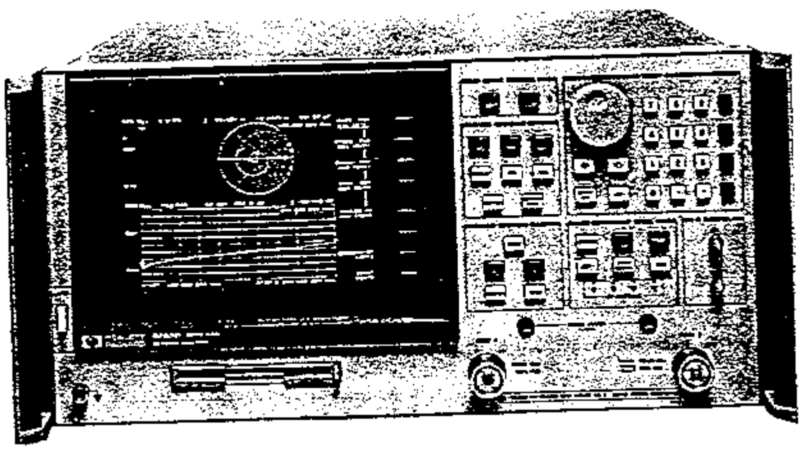
References

TDR Fundamentals, Hewlett Packard App. Note 62

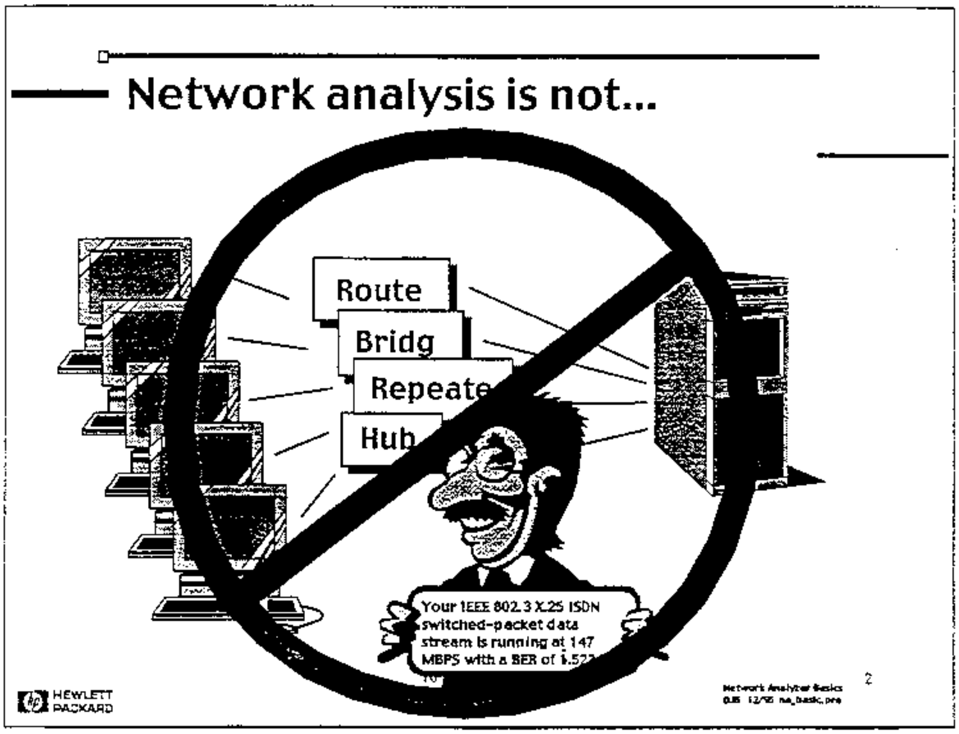
HP teaching tools WWW page, http://www.tmo.hp.com/tmo/iia/edcorner

High Bandwidth Oscilloscope Sampling Architectures, HP product note 54120-3, August 1989

Microwave Measurements with a Vector Network Analyzer



Ralph J. Pasquinelli
Fermilab



This module is not about computer networks! When the name "network analyzer" was coined many years ago, there were no such things as computer networks. Back then, networks always referred to *electrical* networks. Today, when we refer to the things that network analyzers measure, we speak mostly about devices and components.

Reflection Parameters

Reflection Coefficient $\Gamma = \dfrac{V_{\text{reflected}}}{V_{\text{incident}}} = \rho \angle \Phi = \dfrac{Z_L - Z_O}{Z_L + Z_O}$

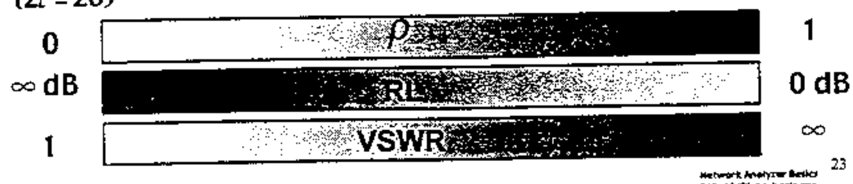
Return loss = -20 log(ρ), $\rho = |\Gamma|$

Voltage Standing Wave Ratio

$$\text{VSWR} = \frac{E_{max}}{E_{min}} = \frac{1+\rho}{1-\rho}$$

***No reflection* (Z_L = Zo)**

***Full reflection* (Z_L = open, short)**



23

Network Analyzer Basics

HEWLETT PACKARD

The first term for reflected waves is reflection coefficient gamma (Γ). The magnitude portion of gamma is called rho (ρ). Reflection coefficient is the ratio of the reflected signal voltage to the incident signal voltage. For example, a transmission line terminated in Zo will have all energy transferred to the load; hence Vrefl = 0 and $\rho = 0$. When ZL is not equal to Zo , some energy is reflected and ρ is greater than zero. When ZL = a short or open circuit, all energy is reflected and $\rho = 1$. The range of possible values for ρ is then zero to one.

Since it is often very convenient to show reflection on a logarithmic display, the second way to convey reflection is return loss. Return loss is expressed in terms of dB, and is a scalar quantity. The definition for return loss includes a negative sign so that the return loss value is always a positive number (when measuring reflection on a network analyzer with a log magnitude format, ignoring the minus sign gives the results in terms of return loss). Return loss can be thought of as the number of dB that the reflected signal is below the incident signal. Return loss varies between infinity for a Zo impedance and 0 dB for an open or short circuit.

As we have already seen, two waves traveling in opposite directions on the same media cause a "standing wave". This condition can be measured in terms of the voltage standing wave ratio (VSWR or SWR for short), and is defined as the maximum value of the RF envelope over the minimum value of the envelope. This value can be computed as $(1+\rho)/(1-\rho)$. VSWR can take on values between one and infinity.

Transmission Parameters

Transmission Coefficient $T = \dfrac{V_{Transmitted}}{V_{Incident}} = \tau\angle\phi$

Insertion Loss (dB) $= -20$ *Log* $\left|\dfrac{V_{Trans}}{V_{Inc}}\right| = -20 \log \tau$

Gain (dB) $= 20$ *Log* $\left|\dfrac{V_{Trans}}{V_{Inc}}\right| = 20 \log \tau$

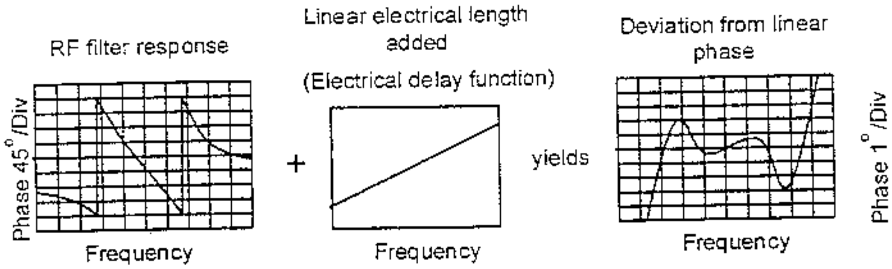
HEWLETT PACKARD

Network Analyzer Basics
DJB 12/95 na_basic.pre
24

Transmission coefficient T is defined as the transmitted voltage divided by the incident voltage. If $|V_{trans}| > |V_{inc}|$, we have gain, and if $|V_{trans}| < |V_{inc}|$, we have attenuation or insertion loss. When insertion loss is expressed in dB, a negative sign is added in the definition so that the loss value is expressed as a positive number. The phase portion of the transmission coefficient is called insertion phase.

Deviation from Linear Phase

Use electrical delay to remove linear portion of phase response

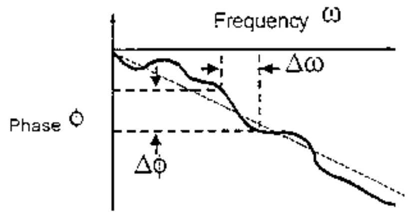


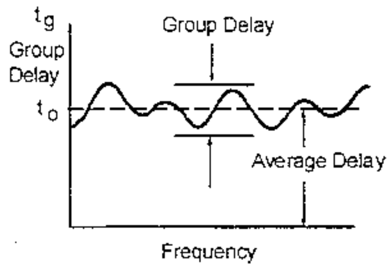
Low resolution

High resolution

Looking at insertion phase directly is usually not very useful. This is because the phase has a negative slope with respect to frequency due to the electrical length of the device (the longer the device, the greater the slope). Since it is only the deviation from linear phase which causes distortion, it is desirable to remove the linear portion of the phase response. This can be accomplished by using the electrical delay feature of the network analyzer to cancel the electrical length of the DUT. This results in a high-resolution display of phase distortion (deviation from linear phase).

What is group delay?





$$\text{Group Delay } (t_g) = \frac{-d\,\phi}{d\,\omega} = \frac{-1}{360^\circ} * \frac{d\,\phi}{d\,f}$$

ϕ in radians
ω in radians/sec
ϕ in degrees
f in Hz ($\omega = 2\pi f$)

Deviation from constant group delay indicates distortion

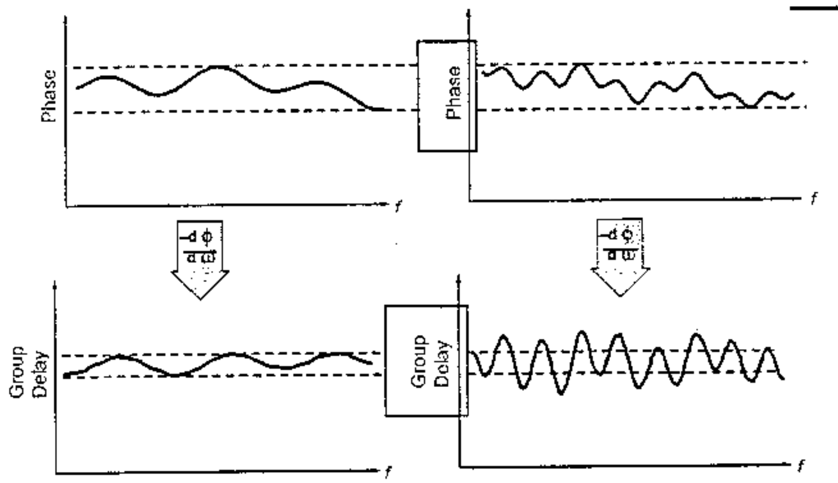
Average delay indicates transit time

HEWLETT PACKARD

Network Analyzer Basics 25

Another useful measure of phase distortion is group delay. Group delay is a measure of the transit time of a signal through the device under test, versus frequency. Group delay is calculated by differentiating the insertion-phase response of the DUT versus frequency. Another way to say this is that group delay is a measure of the slope of the transmission phase response. The linear portion of the phase response is converted to a constant value (representing the average signal-transit time) and deviations from linear phase are transformed into deviations from constant group delay. The variations in group delay cause signal distortion, just as deviations from linear phase cause distortion. Group delay is just another way to look at linear phase distortion.

Why measure group delay?



Same p-p phase ripple can result in different group delay

HEWLETT PACKARD

Why are both deviation from linear phase and group delay commonly measured? Depending on the device, both may be important. Specifying a maximum peak-to-peak value of phase ripple is not sufficient to completely characterize a device since the slope of the phase ripple is dependent on the number of ripples which occur per unit of frequency. Group delay takes this into account since it is the differentiated phase response. Group delay is often a more accurate indication of phase distortion. The plot above shows that the same value of peak-to-peak phase ripple can result in substantially different group delay responses. The response on the right with the larger group-delay variation would cause more signal distortion.

Low-Frequency Network Characterization

H-parameters	*Y-parameters*	*Z-parameters*
$V_1 = h_{11}I_1 + h_{12}V_2$	$I_1 = y_{11}V_1 + y_{12}V_2$	$V_1 = z_{11}I_1 + z_{12}I_2$
$V_2 = h_{21}I_1 + h_{22}V_2$	$I_2 = y_{21}V_1 + y_{22}V_2$	$V_2 = z_{21}I_1 + z_{22}I_2$

$$h_{11} = \left.\frac{V_1}{I_1}\right|_{V_2=0}$$ *(requires short circuit)*

$$h_{12} = \left.\frac{V_1}{V_2}\right|_{I_1=0}$$ *(requires open circuit)*

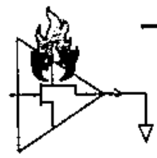
All of these parameters require measuring voltage and current (as a function of frequency)

HEWLETT PACKARD

Network Analyzer Basics DJB 12/96 na_basic.pre 28

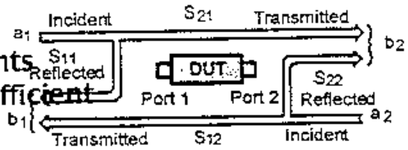
In order to completely characterize an unknown linear two-port device, we must make measurements under various conditions and compute a set of parameters. These parameters can be used to completely describe the electrical behavior of our device (or "network"), even under source and load conditions other than when we made our measurements. For low-frequency characterization of devices, the three most commonly measured parameters are the H, Y and Z-parameters. All of these parameters require measuring the total voltage or current as a function of frequency at the input or output nodes (ports) of the device. Furthermore, we have to apply either open or short circuits as part of the measurement. Extending measurements of these parameters to high frequencies is not very practical.

Limitations of H, Y, Z Parameters
(Why use S-parameters?)



H,Y, Z parameters
- Hard to measure total voltage and current at device ports at high frequencies
- Active devices may oscillate or self-destruct with shorts / opens

S-parameters
- Relate to familiar measurements (gain, loss, reflection coefficient ...)
- Relatively easy to measure
- Can cascade S-parameters of multiple devices to predict system performance
- Analytically convenient
  - CAD programs
  - Flow-graph analysis
- Can compute H, Y, or Z parameters from S-parameters if desired



$b_1 = S_{11} a_1 + S_{12} a_2$

$b_2 = S_{21} a_1 + S_{22} a_2$

HEWLETT PACKARD

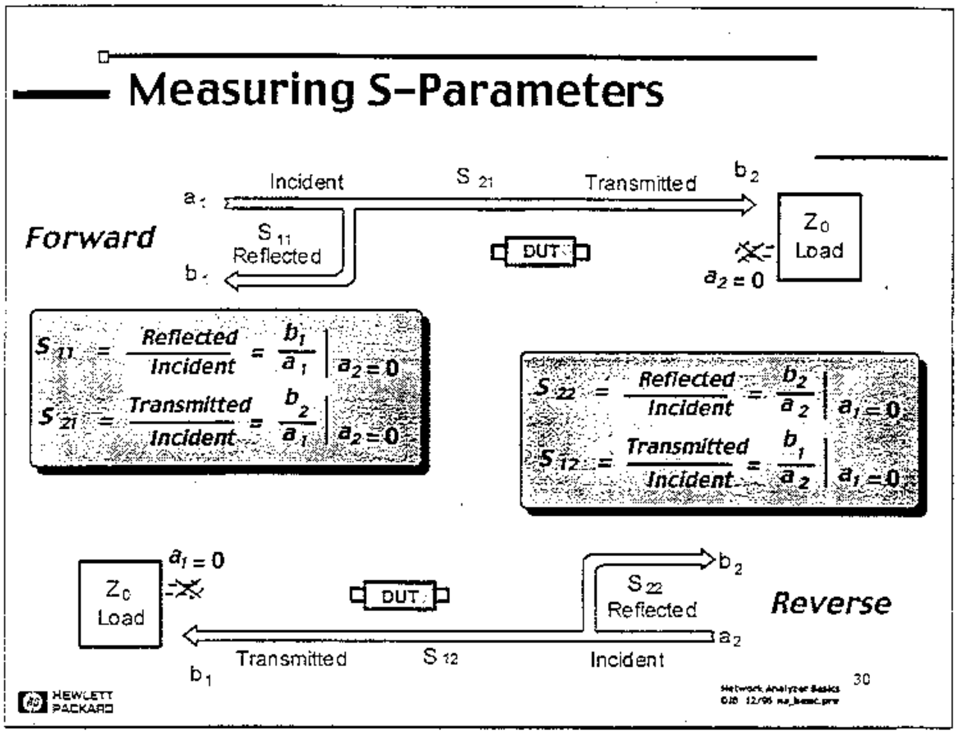
Network Analyzer Basics
DJS 12/95 na_basic.pre

29

At high freq[illegible]ure total voltage and current at the device ports. One cannot simply connect a voltmeter or current probe and get accurate measurements due to the impedance of the probes themselves and the difficulty of placing the probes at the desired positions. In addition, active devices may oscillate or self-destruct with the connection of shorts and opens.

Clearly, some other way of characterizing high-frequency networks is needed that doesn't have these drawbacks. That is why scattering or S-parameters were developed. S-parameters have many advantages over the previously mentioned H, Y or Z-parameters. They relate to familiar measurements such as gain, loss, and reflection coefficient. They are relatively easy to measure, and don't require the connection of undesirable loads to the device under test. The measured S-parameters of multiple devices can be cascaded to predict overall system performance. They are analytically convenient for CAD programs and flow-graph analysis. And, H, Y, or Z-parameters can be derived from S-parameters if desired.

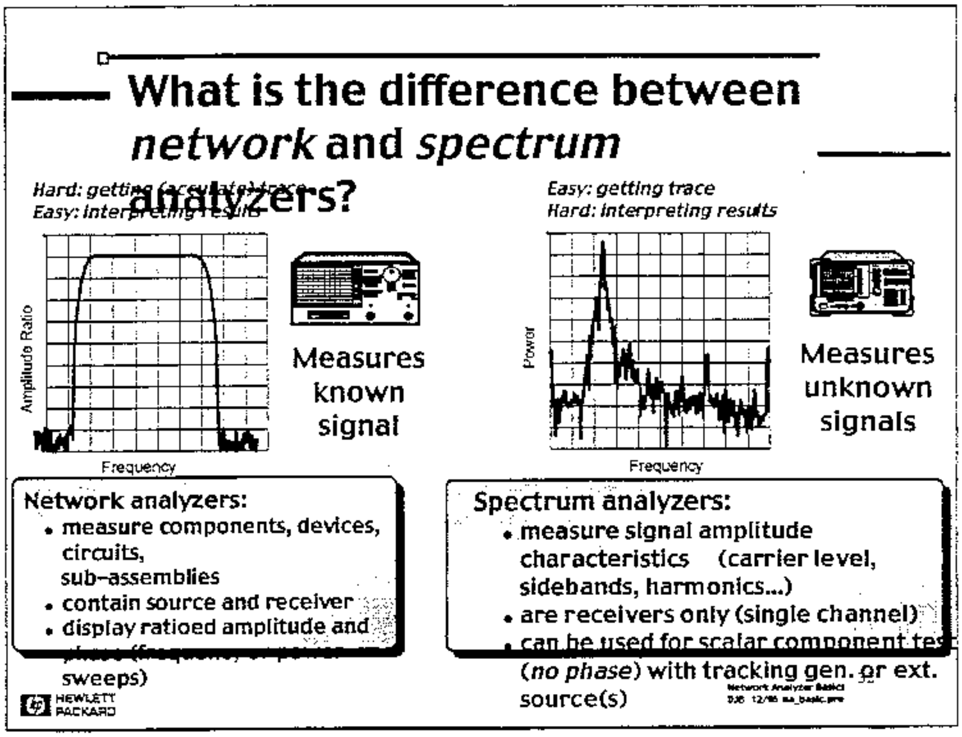
An N-port DUT has N^2 S-parameters. So, a two-port device has four S-parameters. The numbering convention for S-parameters is that the first number following the "S" is the port where energy emerges, and the second number is the port where energy enters. So, S21 is a measure of power coming out port 2 as a result of applying an RF stimulus to port 1. When the numbers are the same (e.g., S11), it indicates a reflection measurement.



S11 and S21 are determined by measuring the magnitude and phase of the incident, reflected and transmitted signals when the output is terminated in a perfect Zo (a load that equals the characteristic impedance of the test system). This condition guarantees that a_2 is zero. S11 is equivalent to the input complex reflection coefficient or impedance of the DUT, and S21 is the forward complex transmission coefficient.

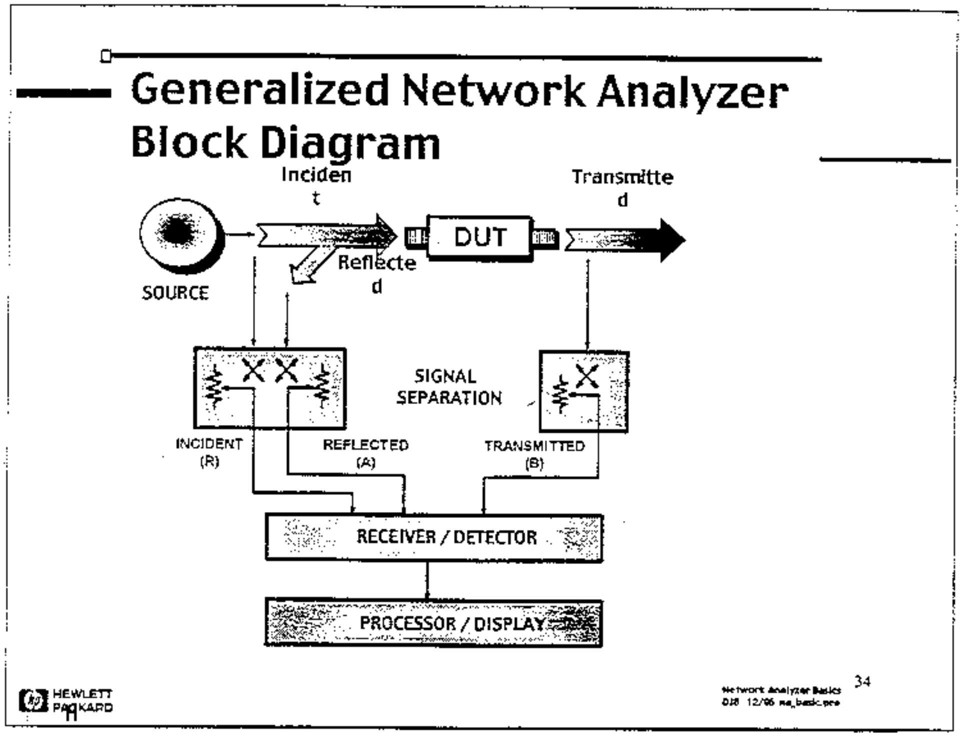
Likewise, by placing the source at port 2 and terminating port 1 in a perfect load (making a_1 zero), S22 and S12 measurements can be made. S22 is equivalent to the output complex reflection coefficient or output impedance of the DUT, and S12 is the reverse complex transmission coefficient.

The accuracy of S-parameter measurements depends greatly on how good a termination we apply to the port not being stimulated. Anything other than a perfect load will result in a_1 or a_2 not being zero (which violates the definition for S-parameters). When the DUT is connected to the test ports of a network analyzer and we don't account for imperfect test port match, we have not done a very good job satisfying the condition of a perfect termination. For this reason, two-port error correction, which corrects for source and load match, is very important for accurate S-parameter measurements (two-port correction is covered in the calibration section).



Now that we have seen some of the measurements that are commonly done with network and spectrum analyzers, it might be helpful to review the main differences between these instruments. Network analyzers are used to measure components, devices, circuits, and sub-assemblies. They contain both a source and multiple receivers, and generally display *ratioed* amplitude and phase information (frequency or power sweeps). A network analyzer is always looking at a *known* signal (in terms of frequency), since it is a stimulus-response system. With network analyzers, it is harder to get an (accurate) trace on the display, but very easy to interpret the results. With vector-error correction, network analyzers provide much higher measurement accuracy than spectrum analyzers.

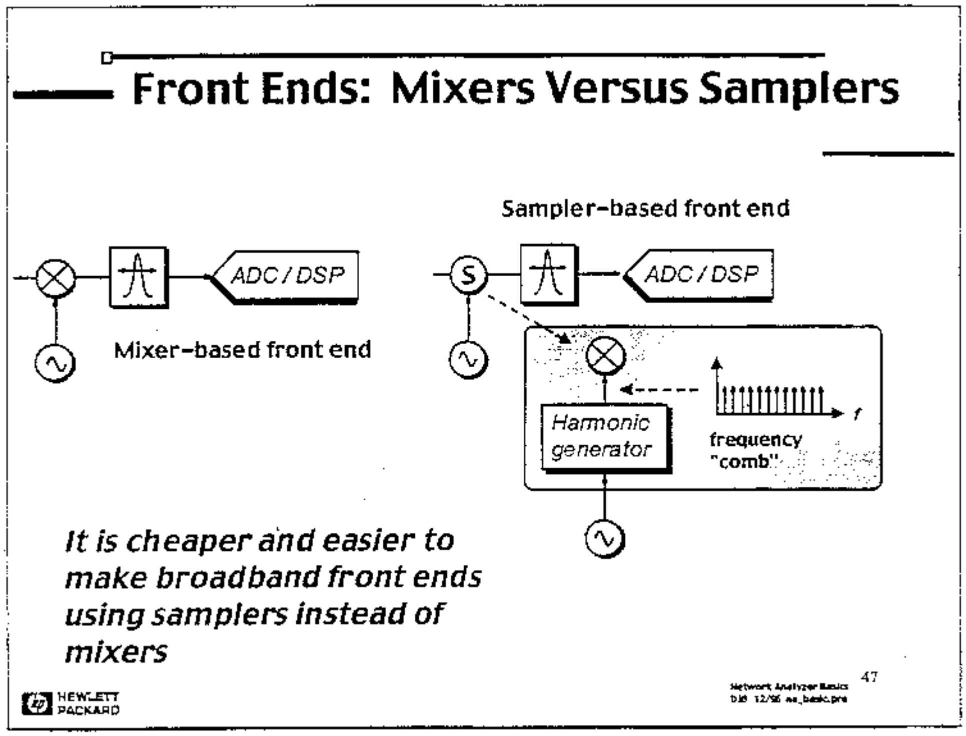
Spectrum analyzers are most often used to measure signal characteristics such as carrier level, sidebands, harmonics, phase noise, etc., on *unknown* signals. They are most commonly configured as a single-channel receiver, without a source. Because of the flexibility needed to analyze signals, spectrum analyzers generally have a much wider range of IF bandwidths available than most network analyzers. Spectrum analyzers are often used with external sources for nonlinear stimulus/response testing. When combined with a tracking generator, spectrum analyzers can be used for scalar component testing (magnitude versus frequency, but no phase measurements). With spectrum analyzers, it is easy to get a trace on the display, but interpreting the results can be much more difficult than with a network analyzer.



Here is a generalized block diagram of a network analyzer, showing the major signal processing sections. In order to measure the incident, reflected and transmitted signal, four sections are required:

1. Source for stimulus
2. Signal-separation devices
3. Receiver that provides detection
4. Processor/display for calculating and reviewing the results

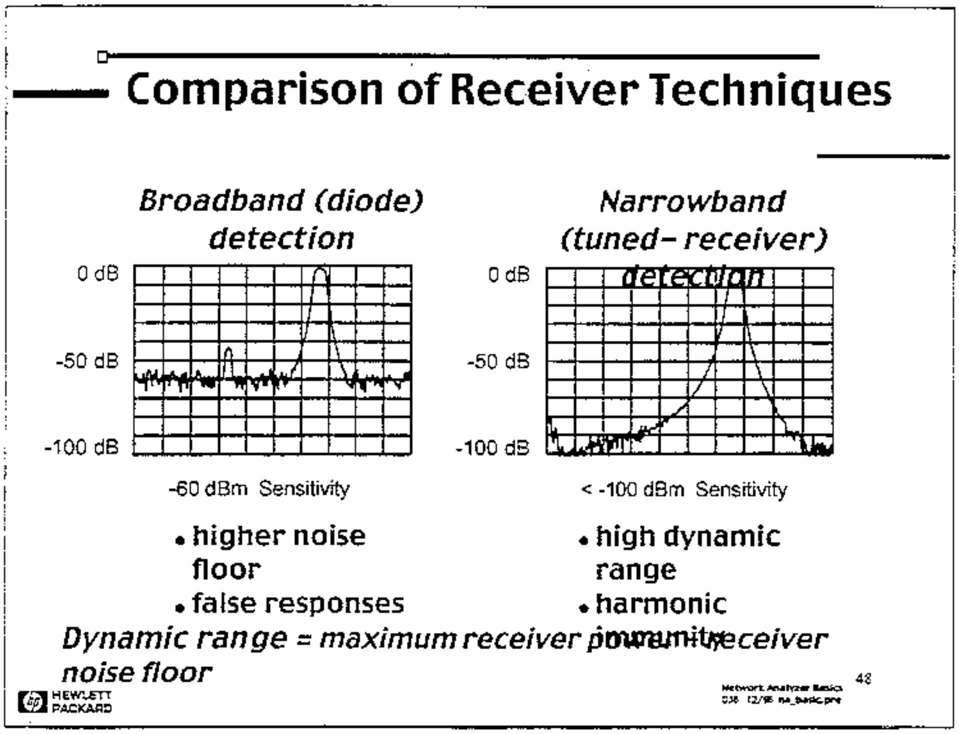
We will examine each of these in more detail.



Tuned receivers can be implemented with mixer- or sampler-based front ends. It is often cheaper and easier to make wideband front ends using samplers instead of mixers, especially for microwave frequency coverage. Samplers are used with many of HP's network analyzers, such as the HP 8753D RF family and the HP 8720D microwave family of analyzers.

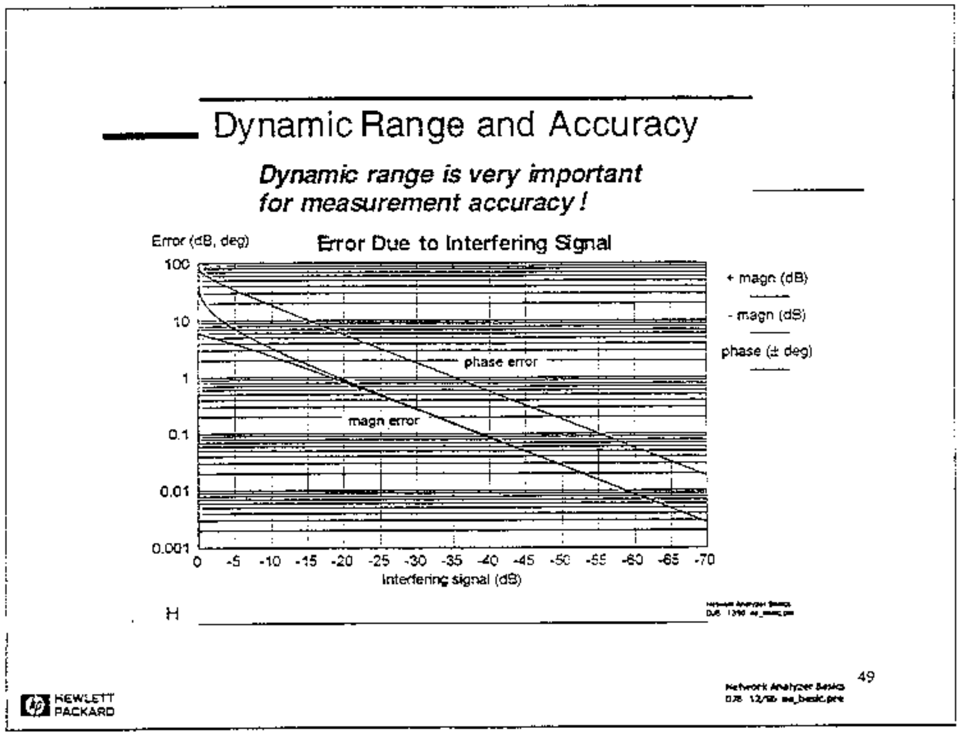
The sampler uses diodes to sample very short time slices of the incoming RF signal. Conceptually, the sampler can be thought of as a mixer with an internal pulse generator. The pulse generator creates a broadband frequency spectrum (often referred to as a "comb") composed of harmonics of the LO. The RF signal mixes with one of the spectral lines (or "comb tooth") to produce the desired IF. Compared to a mixer-based network analyzer, the LO in a sampler-based front end covers a much smaller frequency range, and a broadband mixer is no longer needed. The tradeoff is that the phase-lock algorithms for locking to the various comb teeth is more complex and time consuming.

Sampler-based front ends also have somewhat less dynamic range than those based on mixers and fundamental LOs. This is due to the fact that additional noise is converted into the IF from all of the comb teeth. Network analyzers with narrowband detection based on samplers still have far greater dynamic range than analyzers that use diode detection.



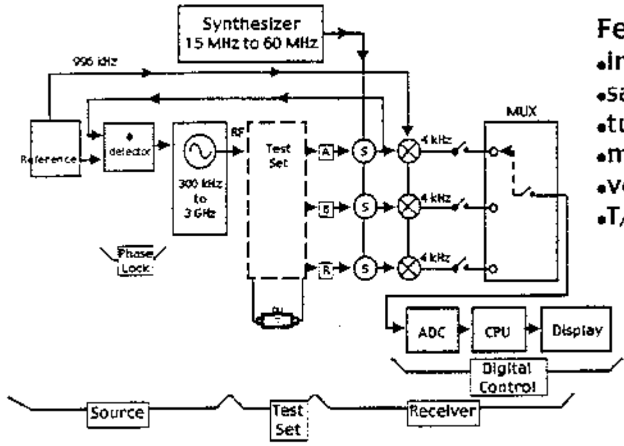
Dynamic range is generally defined as the maximum power the receiver can accurately measure minus the receiver noise floor. There are many applications requiring large dynamic range. One of the most common are filter applications. As you can see here, at least 80 dB dynamic range is needed to properly characterize the rejection characteristics of this filter. The plots show a typical narrowband filter measured on an HP 8757 scalar network analyzer and on the HP 8510 vector network analyzer. Notice that the filter exhibits 90 dB of rejection but the scalar analyzer is unable to measure it because of its higher noise floor.

In the case where the scalar network analyzer was used with broadband diode detection, a harmonic or subharmonic from the source created a "false" response. For example, at some point on a broadband sweep, the second harmonic of the source might fall within the passband of the filter. If this occurs, the detector will register a response, even though the stopband of the filter is severely attenuating the frequency of the fundamental. This response from the second harmonic would show on the display at the frequency of the fundamental. On the tuned receiver, a spurious response such as this would be filtered away and would not appear on the display.

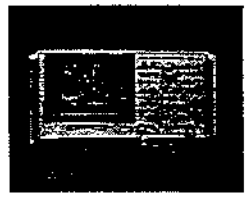


This plot shows the effect that interfering signals (sinusoids or noise) have on measurement accuracy. To get low measurement uncertainty, more dynamic range is needed than the device exhibits. For example, to get less than 0.1 dB magnitude error and less than 1 degree phase error, our noise floor needs to be more than 35 dB below our measured power levels! HP network analyzers often have a clear competitive advantage by providing greater dynamic range than competitor's network analyzers.

Modern Vector Analyzer



Features:
- integrated source
- sampler-based front end
- tuned receiver
- magnitude and phase
- vector-error correction
- T/R or S-parameter test sets



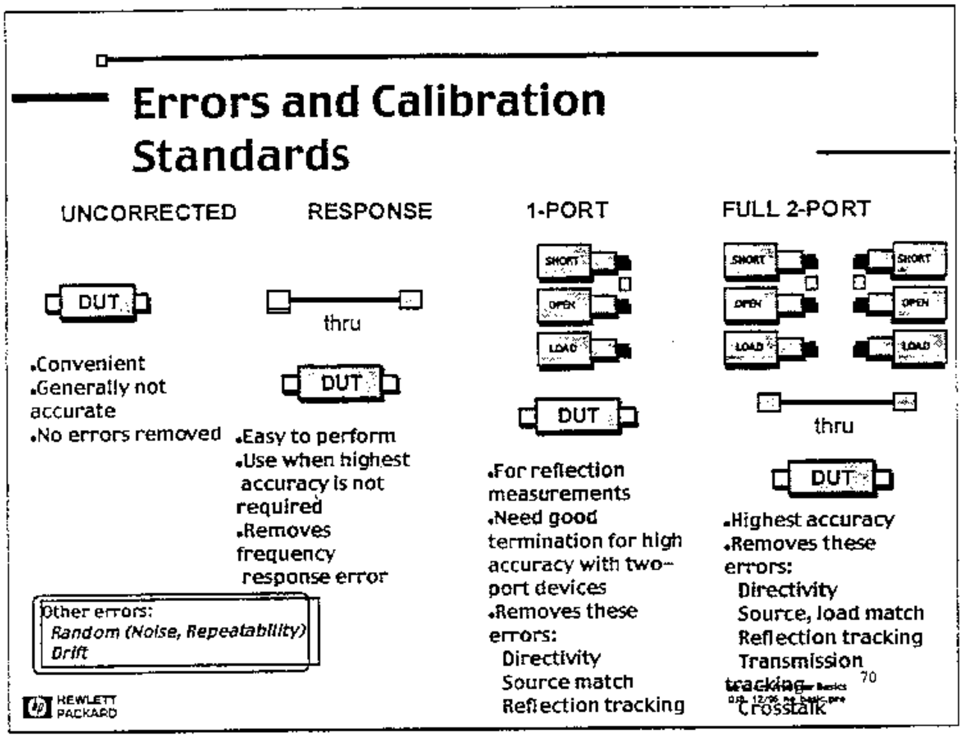
Note: modern scalar analyzers like HP 8711/13C look just like vector analyzers, but they don't display phase

HEWLETT PACKARD

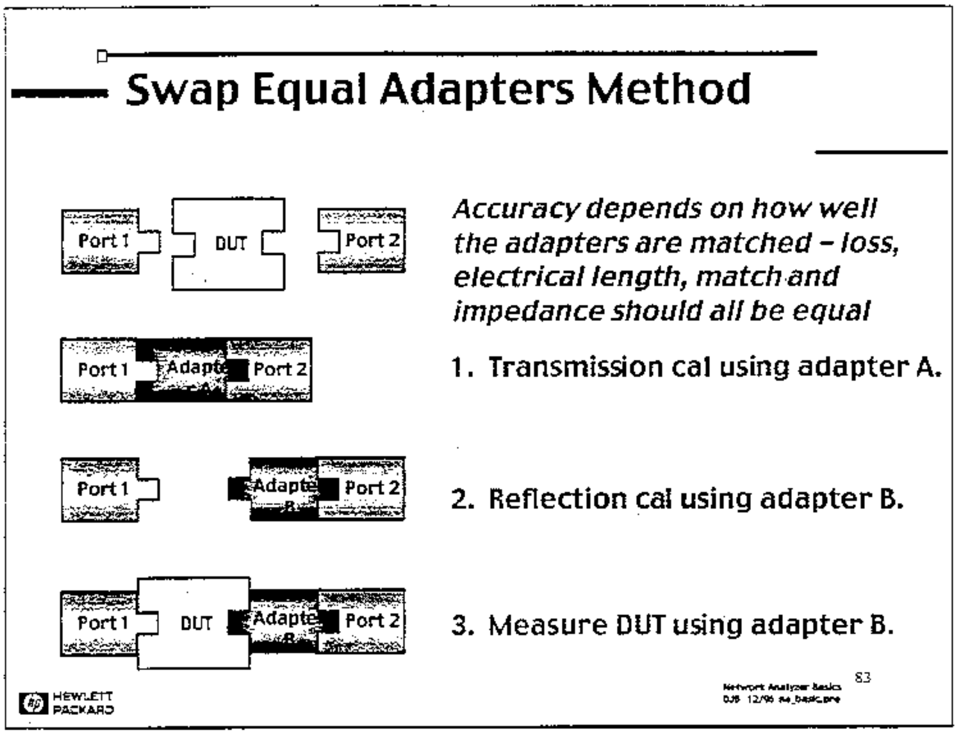
Network Analyzer Basics 53

Here is a block diagram of a modern vector network analyzer. It features an integrated source, a sampler-based front end, and a tuned receiver providing magnitude and phase data with vector-error correction. The test set (the portion of the instrument that contains the signal-separation devices and the switches for directing the RF power) can either be transmission/reflection (T/R) based or an S-parameter test set.

Modern scalar analyzers like the 8711C and 8713C look very similar to this block diagram, but they don't display phase information on the screen. Internally, however, they are essentially vector analyzers. This capability lets them make much more accurate measurements than traditional scalar analyzers.



Here is a summary of the basic error-correction choices available for network analysis measurements.



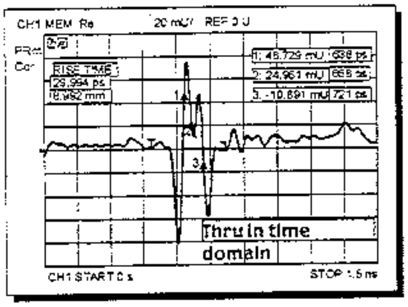
The swap equal adapter method requires the use of two precision matched adapters which are "equal", but have connectors of different sexes (e.g., 7 mm/Type N (m) and 7 mm/Type N (f)) To be equal, the adapters need to have the same match, Zo, insertion loss, and electrical delay. Many of HP's calibration kits include matched adapters.

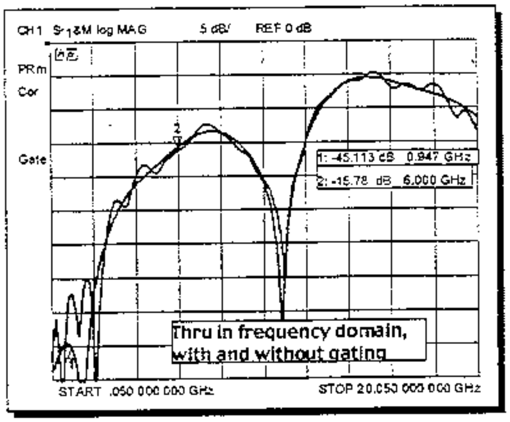
The first step in the procedure is to perform a transmission calibration using the first adapter. Then, adapter A is removed, and adapter B is placed on port two. Adapter B becomes the effective test port. The reflection calibration is then performed on both test ports. Then the DUT is measured with adapter B in place.

The errors remaining after calibration with this method are equal to the differences between the two adapters that are used. The technique provides good accuracy, but not as good as the more complicated adapter-removal technique.

Time-Domain Gating

- TDR and gating can remove undesired reflections
  (a form of error correction)
- Only useful for broadband devices (a load or thru for example)
- Define gate to only include
- Use two-port calibration





92

Network Analyzer Basics

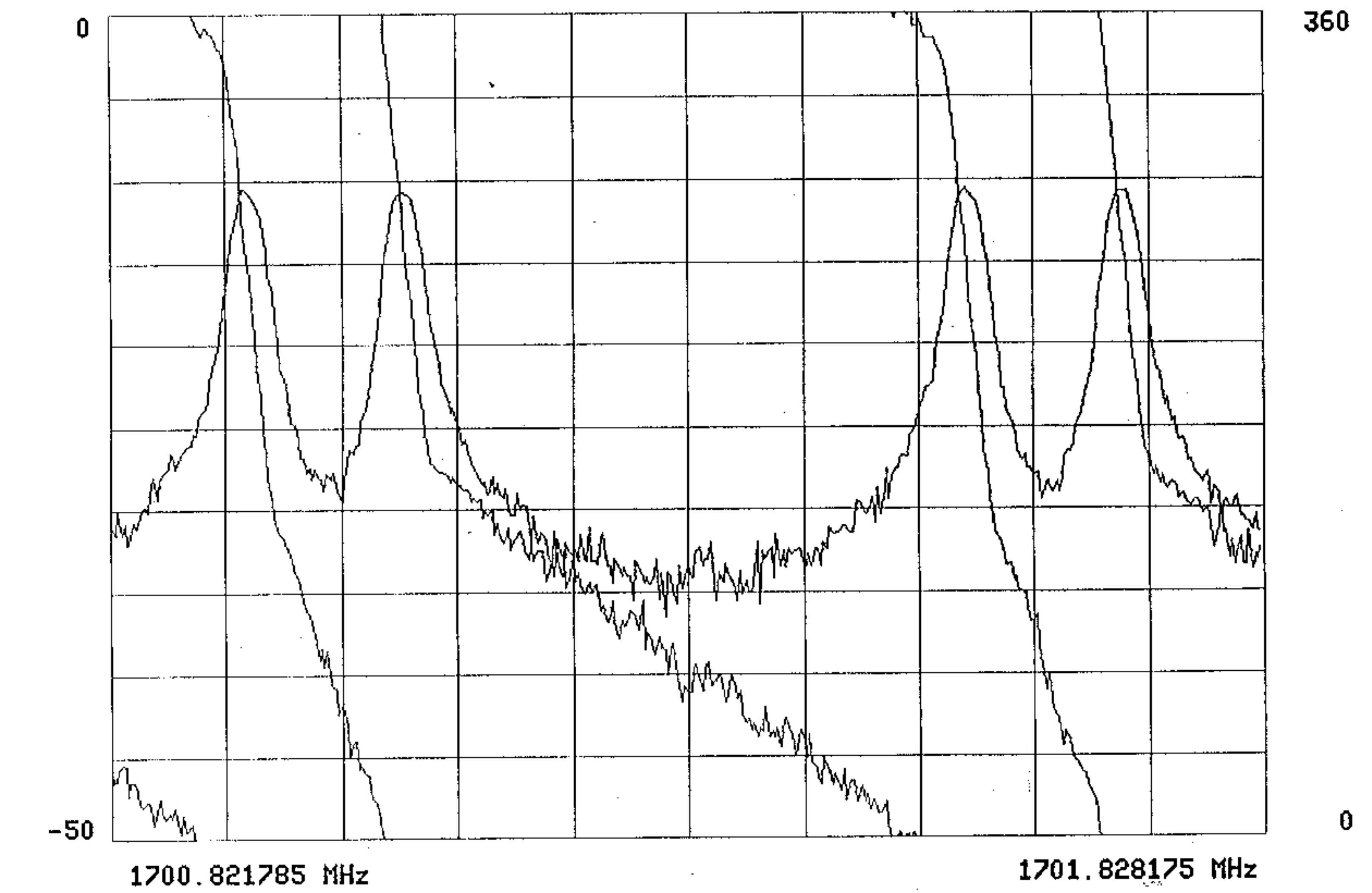
HEWLETT PACKARD

Gating can be used in conjunction with time-domain measurements to separate and remove undesirable reflections from those of interest. For example, gating can isolate the reflections of a DUT in a fixture from those of the fixture itself. This is a form of error correction. For time-domain gating to work effectively, the time domain responses need to be well-separated in time (and therefore distance). The gate itself looks like a filter in time, and has a finite transition range between passing and rejecting a reflection (similar to the skirts of a filter in the frequency domain).

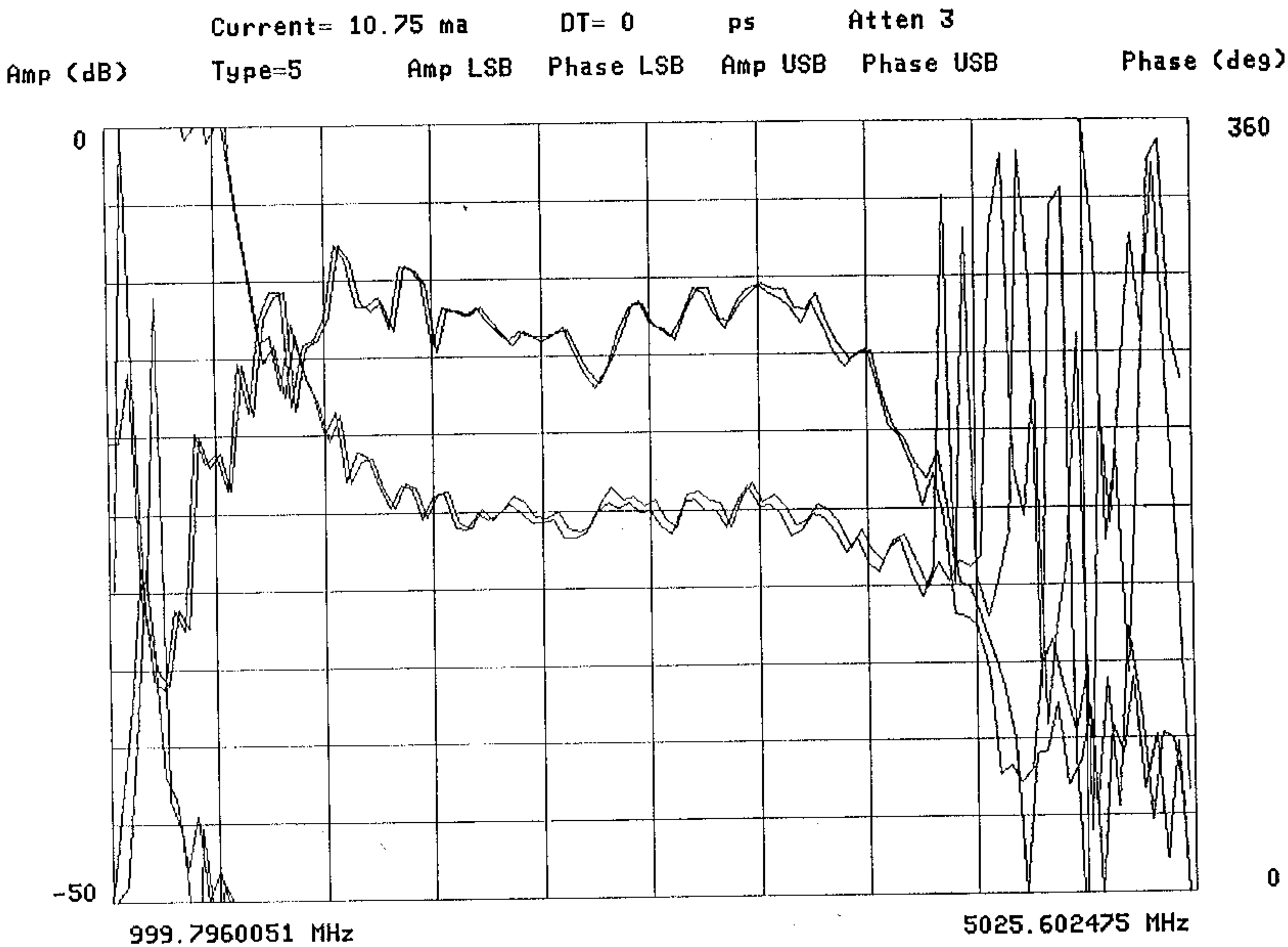
The plots above show the performance of an in-fixture thru standard (without normalization). We see about a 7 dB improvement in return loss at 947 MHz using time-domain gating, resulting in a return loss of 45 dB. The gating effectively removes the effects of the SMA connectors at either end of the test fixture.

File number = 5
Core 2-4 GHz Vert.   NB H=2705   07/01/97 2149

Current= 10.74 ma   DT= 0   ps   Atten 3



File number = 2
Core 2-4 GHz Vert.   WB ATTEN = 3   07/01/97 2117



References

HP teaching tools WWW page, http://www.tmo.hp.com/tmo/iia/edcorner

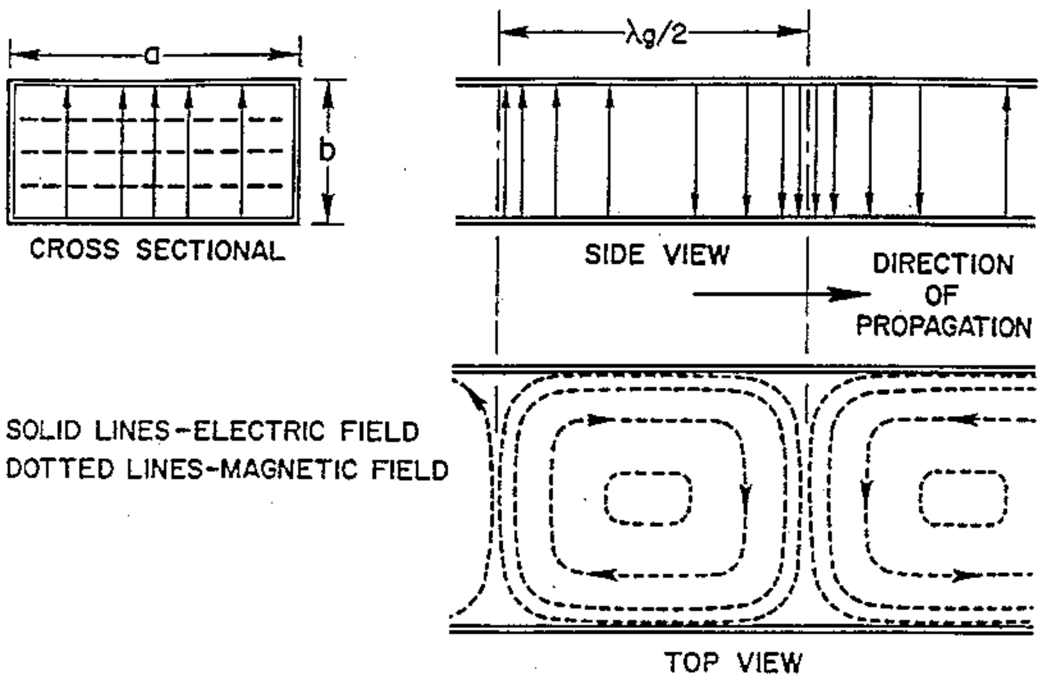
Ralph J. Pasquinelli
Fermilab

Microwave Measurements Laboratory

Components and Devices

Ralph J. Pasquinelli
Fermilab

Rectangular Wave Guide



Advantages: **Low Loss, High Power**

Disadvantages: **Narrow bandwidth, bulky at low frequencies**

$$\lambda g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{\lambda c}\right)^2}}$$

$$\lambda_{C_{TE10}} = 2a$$

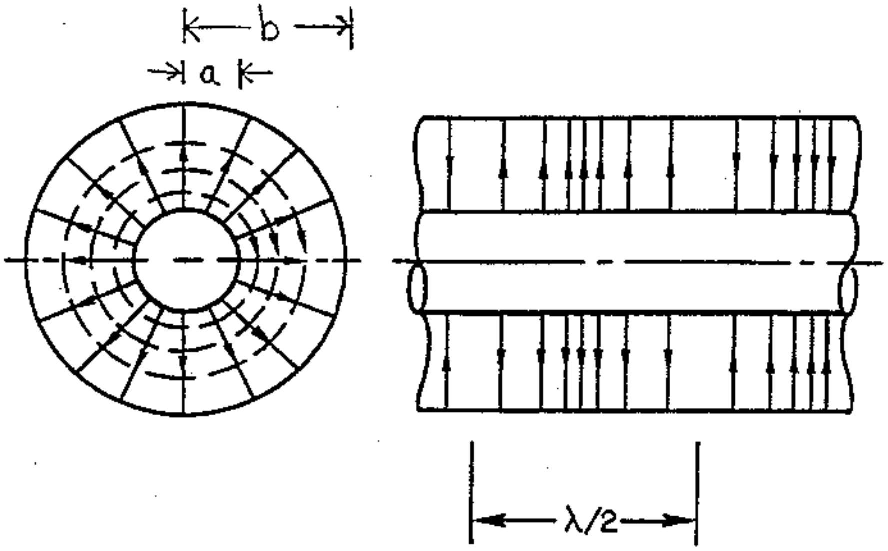
Ralph J. Pasquinelli
Fermilab

Rectangular Wave Characteristics

EIA Waveguide Designation Standard (RS-261-A)	JAN Waveguide Designation (MIL-HDBK-216, 4 January 1962)	Outer Dimensions and Wall Thickness (in inches)	Frequency Range in Gigahertz for Dominant ($TE_{1,0}$) Mode	Cutoff Wavelength λ_c in Centimeters for $TE_{1,0}$ Mode	Cutoff Frequency f_c in Gigahertz for $TE_{1,0}$ Mode	Theoretical Attenuation, Lowest to Highest Frequency in dB/100 ft	Theoretical Power Rating in Megawatts for Lowest to Highes Frequency*
WR-2300	RG-290/U†	23.250×11.750×0.125	0.32–0.49	116.8	0.256	0.051–0.031	153.0–212.0
WR-2100	RG-291/U†	21.250×10.750×0.125	0.35–0.53	106.7	0.281	0.054–0.034	120.0–173.0
WR-1800	RG-201/U†	18.250×9.250×0.125	0.425–0.620	91.4	0.328	0.056–0.038	93.4–131.9
WR-1500	RG-202/U†	15.250×7.750×0.125	0.49–0.740	76.3	0.393	0.069–0.050	67.6–93.3
WR-1150	RG-203/U†	11.750×6.000×0.125	0.64–0.96	58.4	0.514	0.128–0.075	35.0–53.8
WR-975	RG-204/U†	10.000×5.125×0.125	0.75–1.12	49.6	0.605	0.137–0.095	27.0–38.5
WR-770	RG-205/U†	7.950×4.100×0.125	0.96–1.45	39.1	0.767	0.201–0.136	17.2–24.1
WR-650	RG-69/U	6.660×3.410×0.080	1.12–1.70	33.0	0.908	0.317–0.212	11.9–17.2
WR-510	—	5.260×2.710×0.080	1.45–2.20	25.9	1.16	—	—
WR-430	RG-104/U	4.460×2.310×0.080	1.70–2.60	21.8	1.375	0.588–0.385	5.2–7.5
WR-340	RG-112/U	3.560×1.860×0.080	2.20–3.30	17.3	1.735	0.877–0.572	—
WR-284	RG-48/U	3.000×1.500×0.080	2.60–3.95	14.2	2.08	1.102–0.752	2.2–3.2
WR-229	—	2.418×1.273×0.064	3.30–4.90	11.6	2.59	—	—
WR-187	RG-49/U	2.000×1.000×0.064	3.95–5.85	9.50	3.16	2.08–1.44	1.4–2.0
WR-159	—	1.718×0.923×0.064	4.90–7.05	8.09	3.71	—	—
WR-137	RG-50/U	1.500×0.750×0.064	5.85–8.20	6.98	4.29	2.87–2.30	0.56–0.71
WR-112	RG-51/U	1.250×0.625×0.064	7.05–10.00	5.70	5.26	4.12–3.21	0.35–0.46
WR-90	RG-52/U	1.000×0.500×0.050	8.20–12.40	4.57	6.56	6.45–4.48	0.20–0.29
WR-75	—	0.850×0.475×0.050	10.00–15.00	3.81	7.88	—	—
WR-62	RG-91/U	0.702×0.391×0.040	12.40–18.00	3.16	9.49	9.51–8.31	0.12–0.16
WR-51	—	0.590×0.335×0.040	15.00–22.00	2.59	11.6	—	—
WR-42	RG-53/U	0.500×0.250×0.040	18.00–26.50	2.13	14.1	20.7–14.8	0.043–0.058
WR-34	—	0.420×0.250×0.040	22.00–33.00	1.73	17.3	—	—
WR-28	RG-96/U‡	0.360×0.220×0.040	26.50–40.00	1.42	21.1	21.9–15.0	0.022–0.031

Ralph J. Pasquinelli
Fermilab

Coaxial Transmission Lines



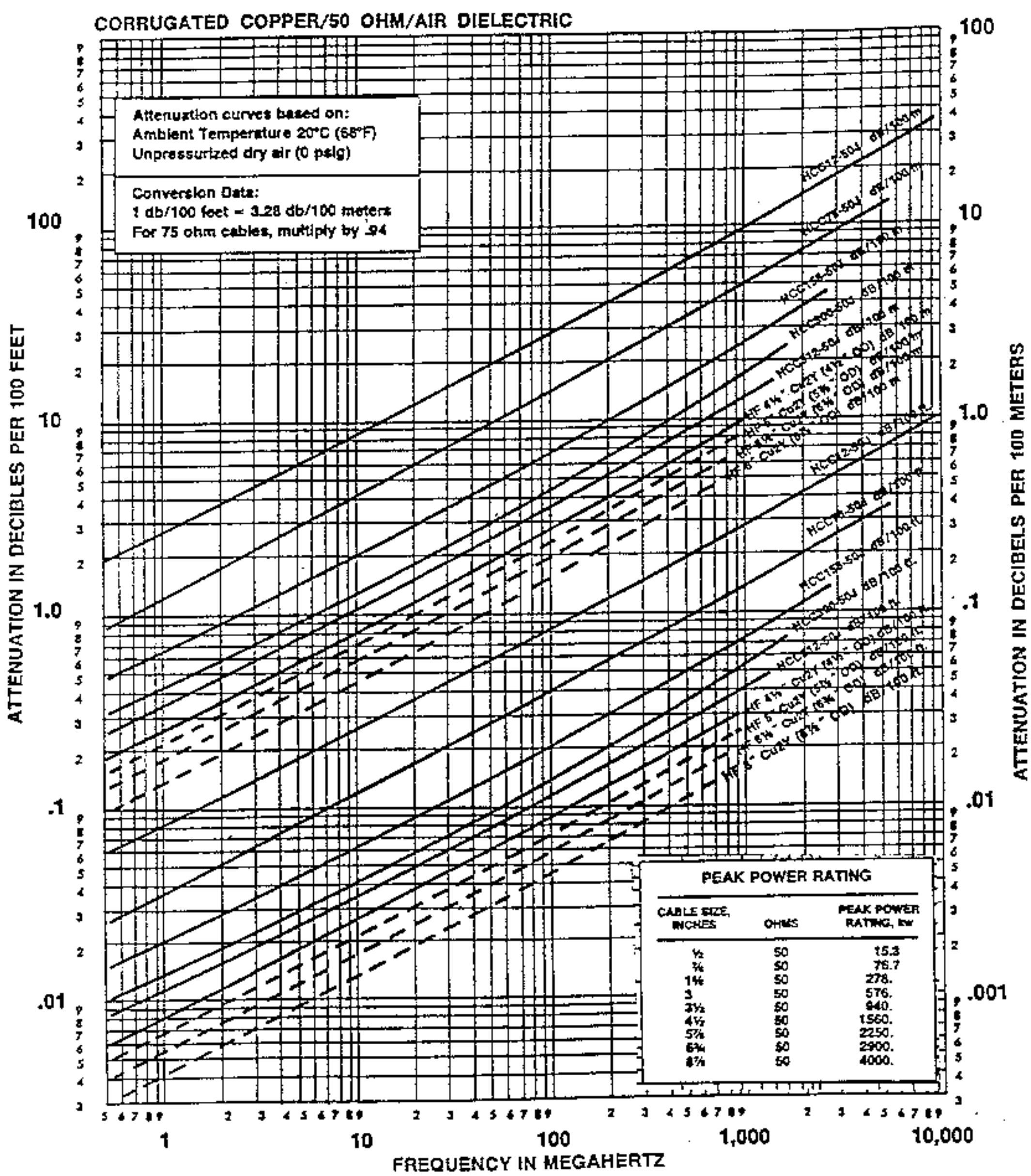
Advantages:	**High Bandwidth, Small size**
Disadvantages:	**High loss, low power**

$$Z_o = 60\sqrt{\frac{\mu_r}{\varepsilon_r}}\ln\frac{b}{a}\,\Omega$$

Ralph J. Pasquinelli
Fermilab

Coax Characteristics

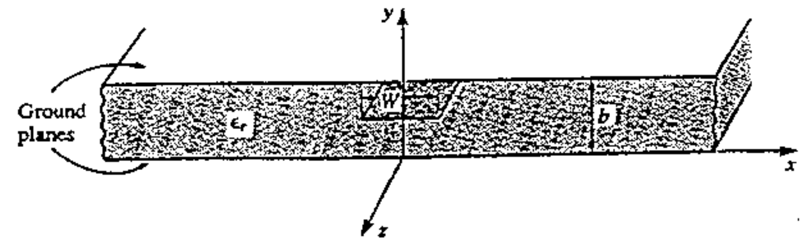
Attenuation — **Air Flexwell Cable**

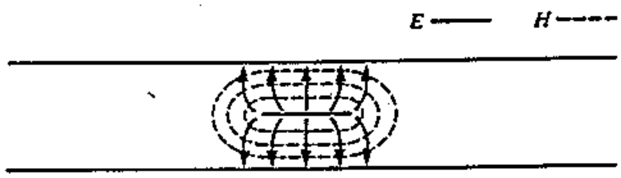


CABLE SIZE, INCHES	OHMS	PEAK POWER RATING, kw
½	50	15.3
⅞	50	76.7
1⅝	50	278.
3	50	576.
3½	50	940.
4½	50	1560.
5⅝	50	2250.
6⅛	50	2900.
8⅞	50	4000.

Ralph J. Pasquinelli
Fermilab

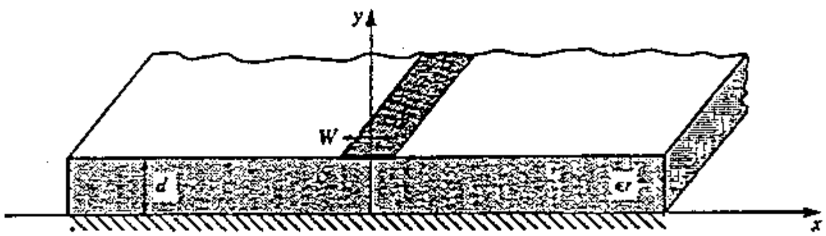
Stripline

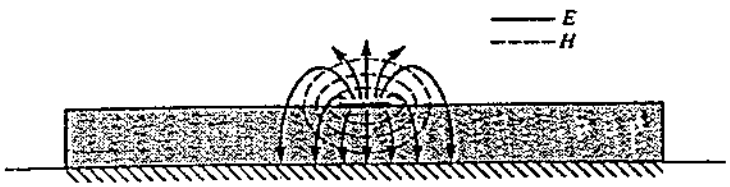




USED for
HYBRIDS
COUPLERS
FILTERS

Microstrip

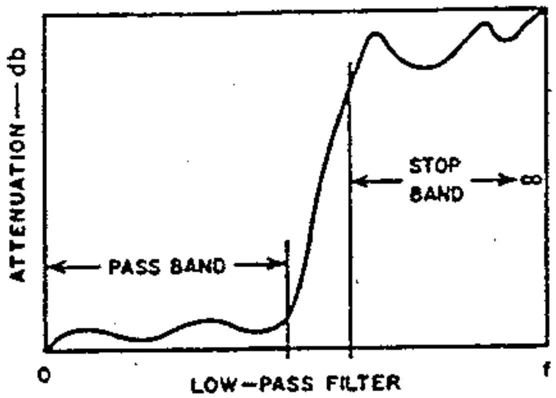




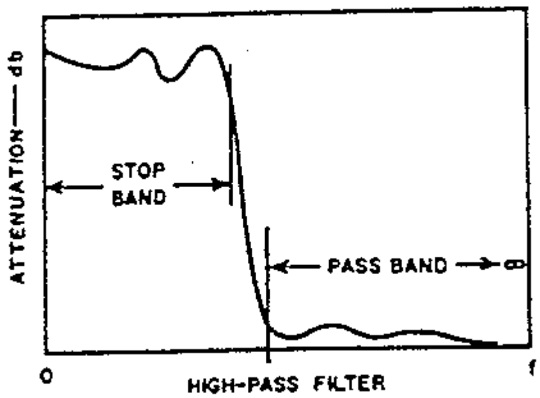
USED for
AMPLIFIERS
MIXERS
FILTERS

Ralph J. Pasquinelli
Fermilab

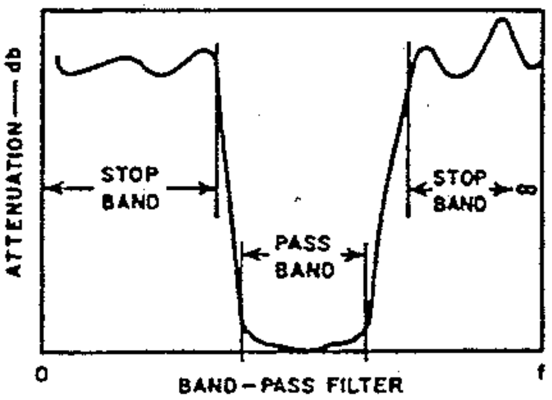
Filters



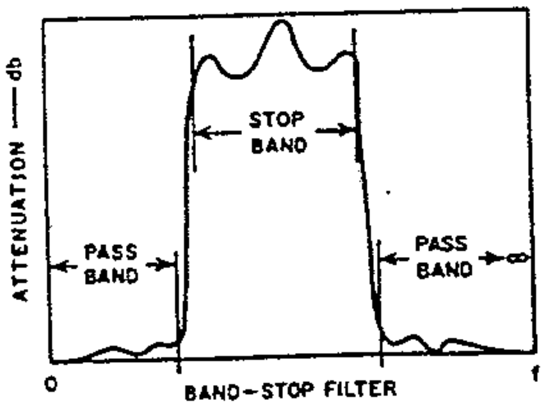
Lowpass



Highpass



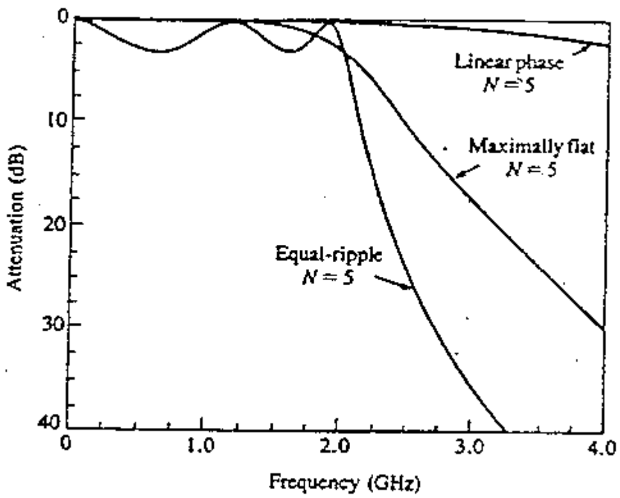
Bandpass

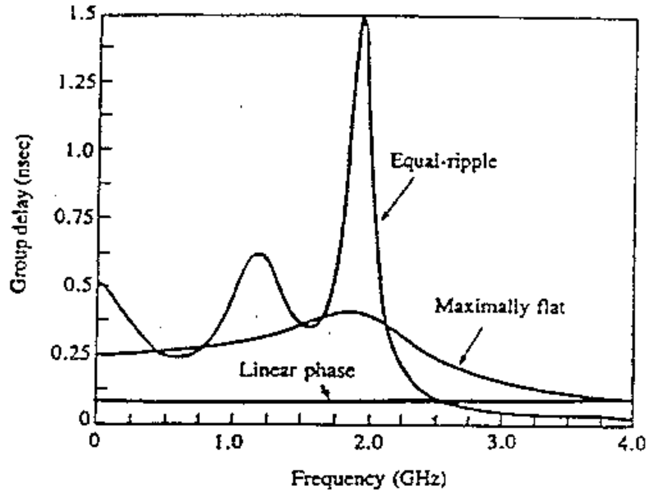


Bandstop

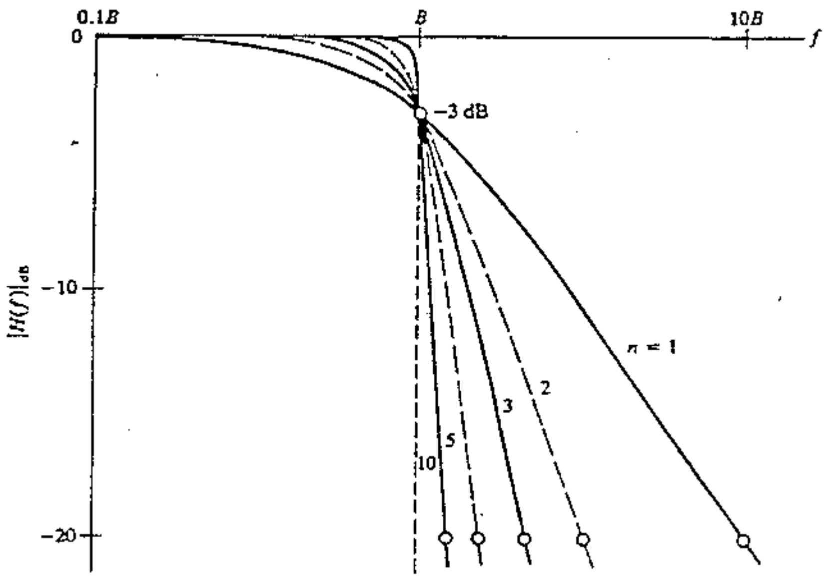
Ralph J. Pasquinelli
Fermilab

Filters Transfer Functions



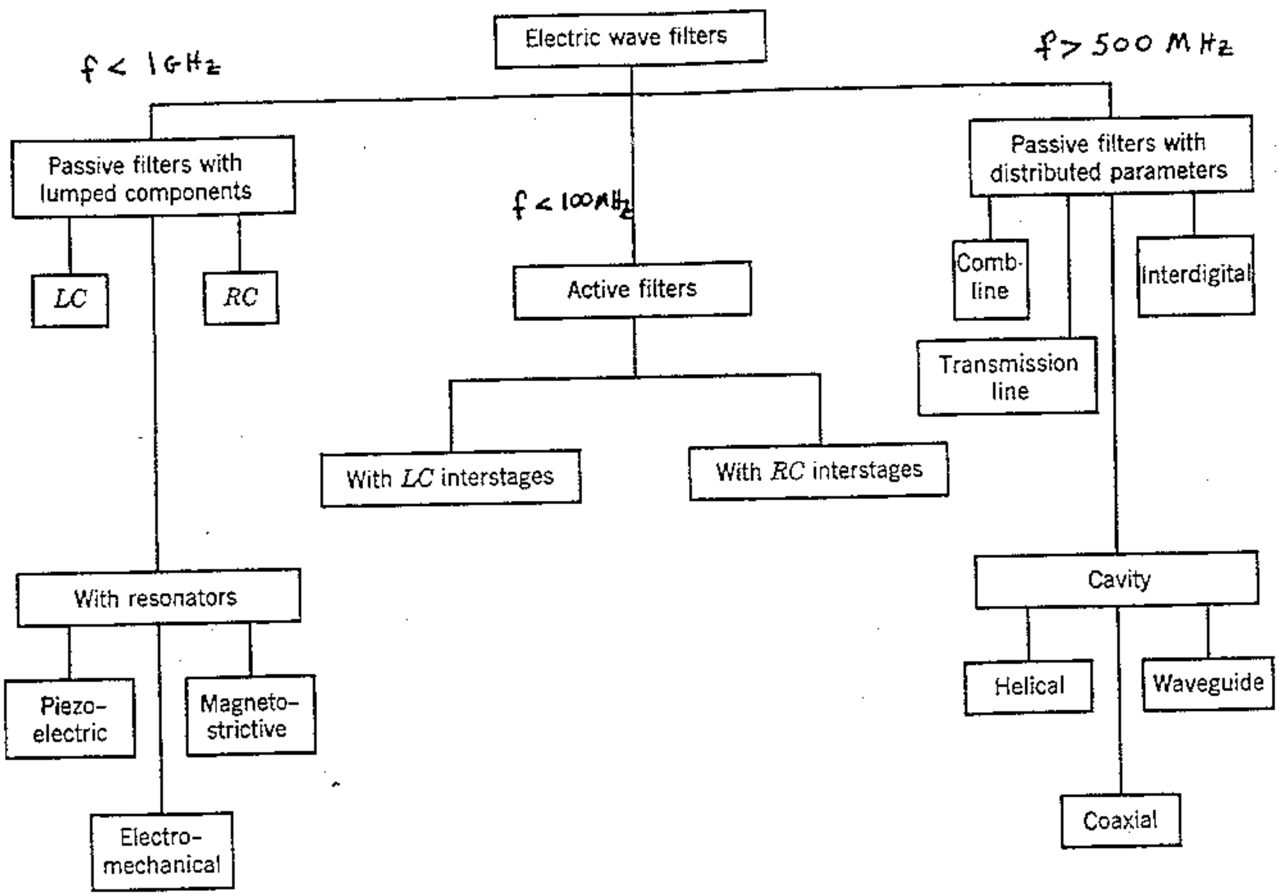


Transfer function for Butterworth lowpass filters of order n:



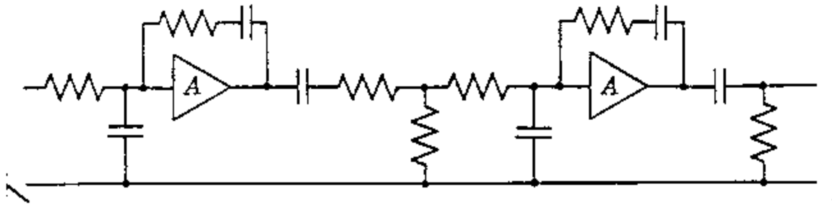
Ralph J. Pasquinelli
Fermilab

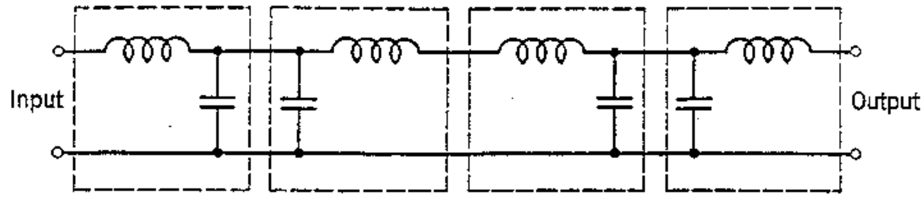
Filter Classifications

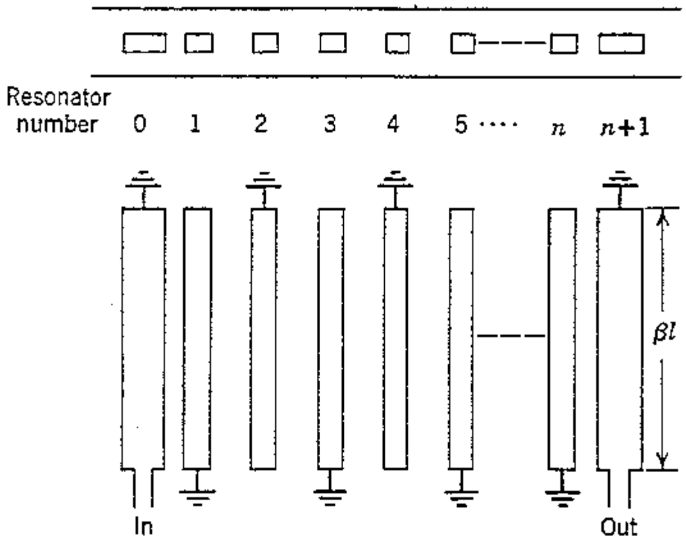


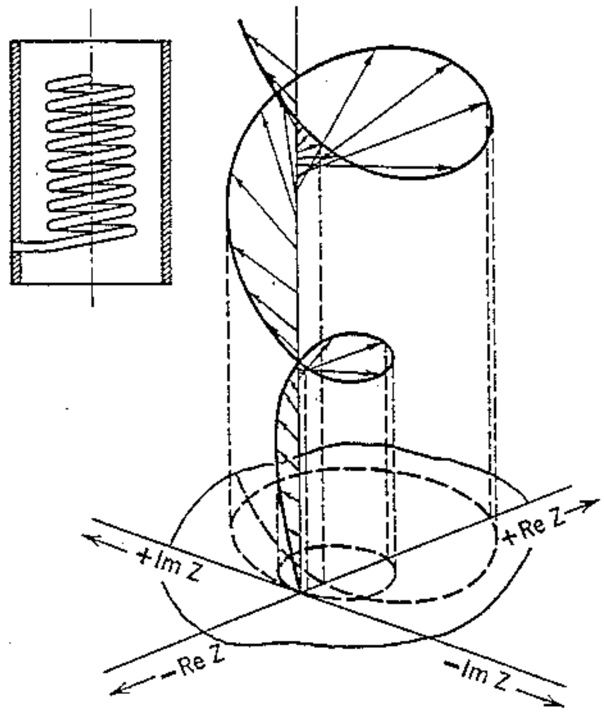
Ralph J. Pasquinelli
Fermilab

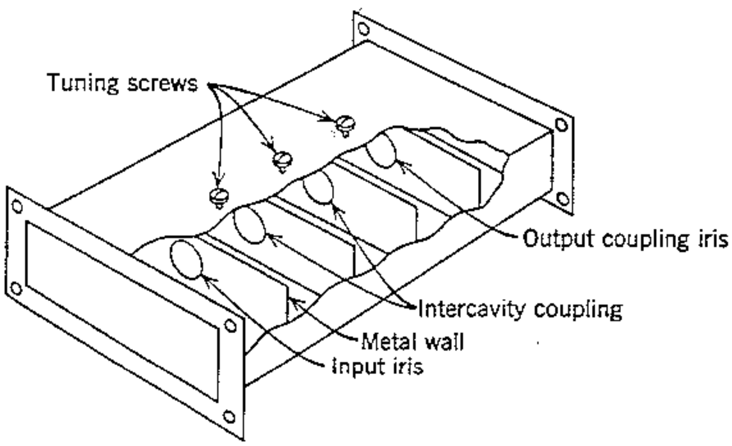
Filter Topologies

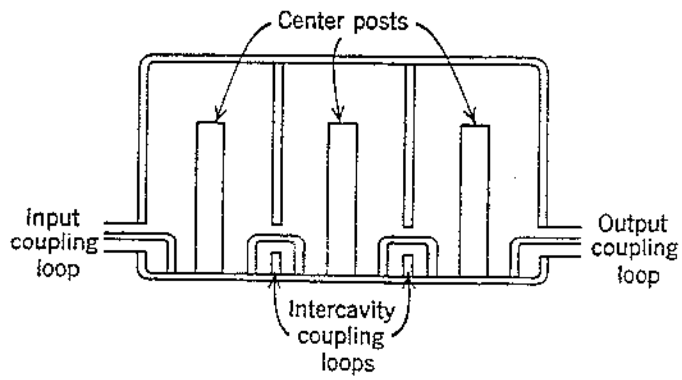






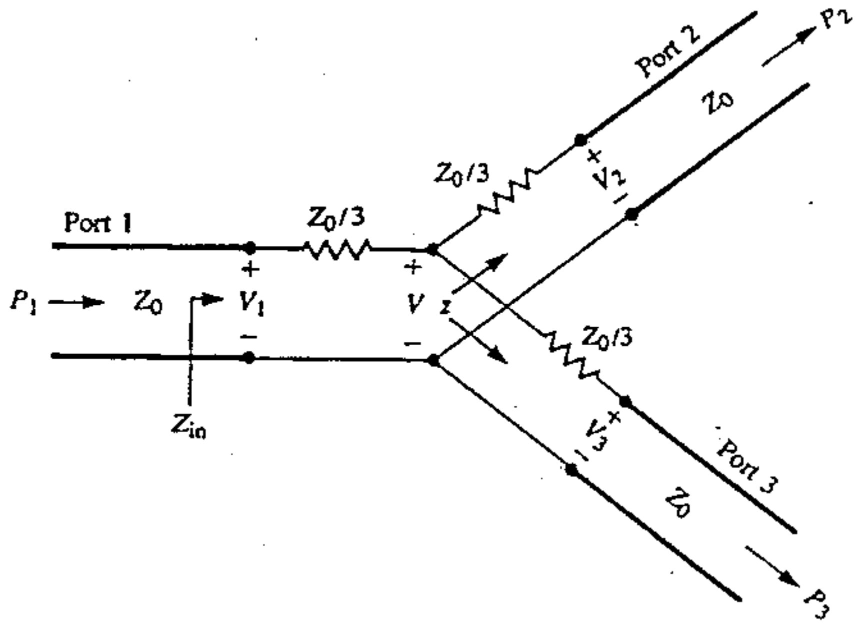






Ralph J. Pasquinelli
Fermilab

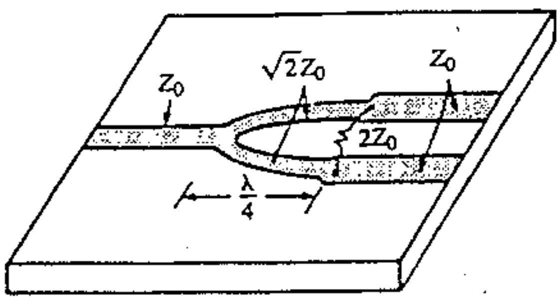
Resistive Power Splitter/Combiner

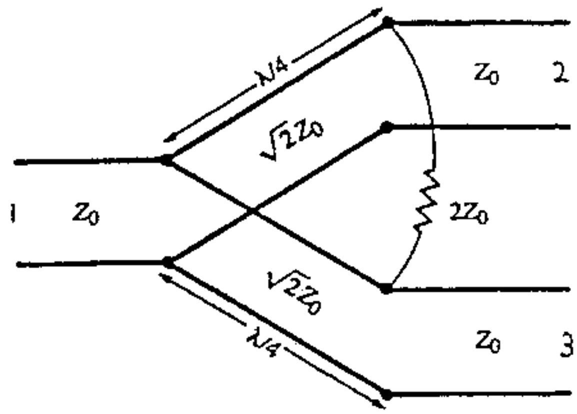


Advantages: **Multi Octave Broad Band**
Disadvantages: **High Loss, low isolation**

Ralph J. Pasquinelli
Fermilab

Wilkinson Power Splitter/Combiner



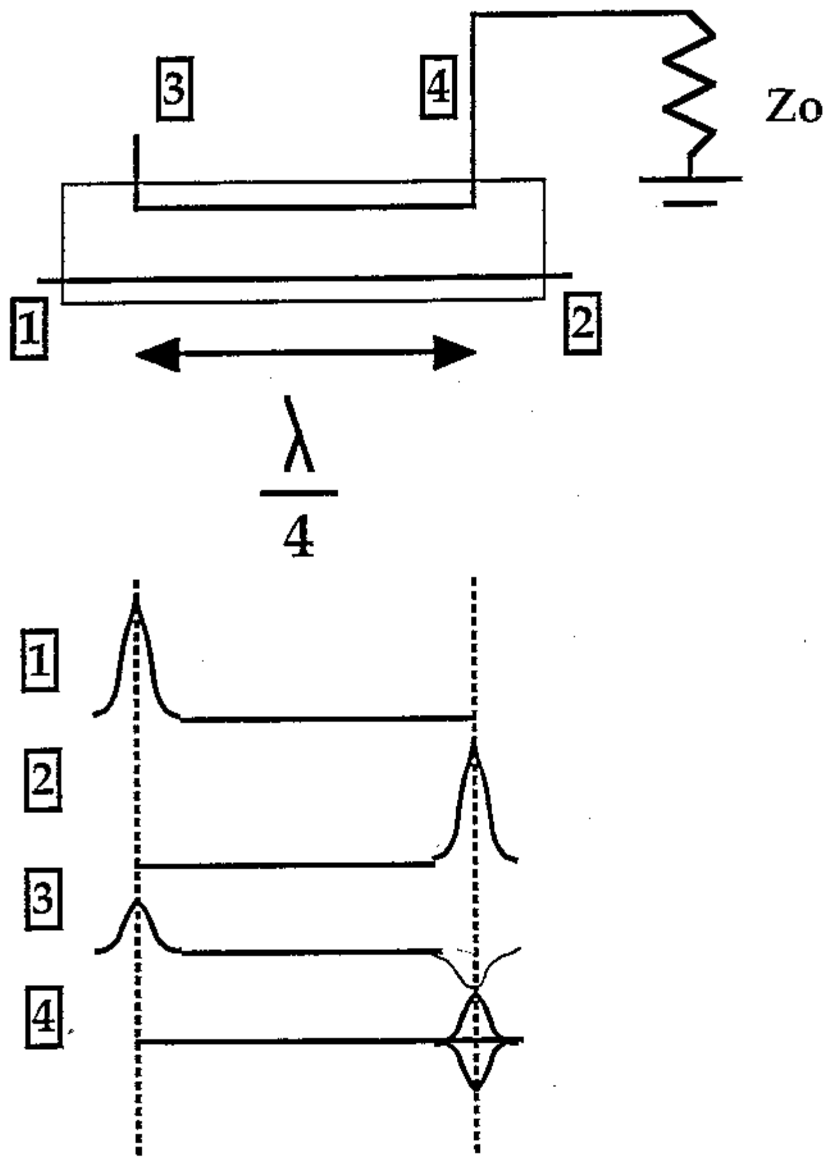


$$\text{S parameters} = \begin{bmatrix} 0 & \frac{-j}{\sqrt{2}} & \frac{-j}{\sqrt{2}} \\ \frac{-j}{\sqrt{2}} & 0 & 0 \\ \frac{-j}{\sqrt{2}} & 0 & 0 \end{bmatrix}$$

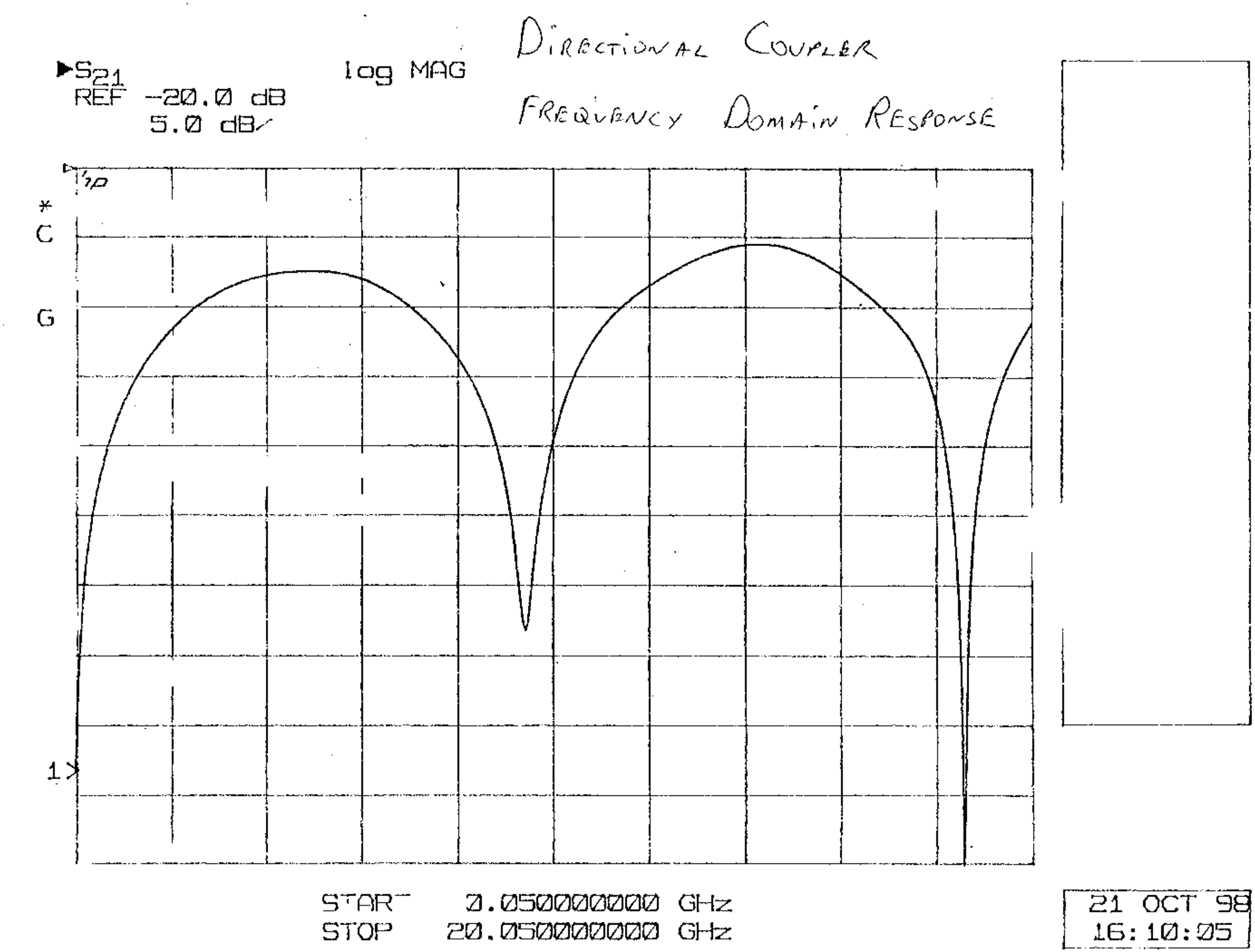
Advantages:	**All ports matched and isolated, low insertion loss**
Disadvantages:	**Octave bandwidths typical**

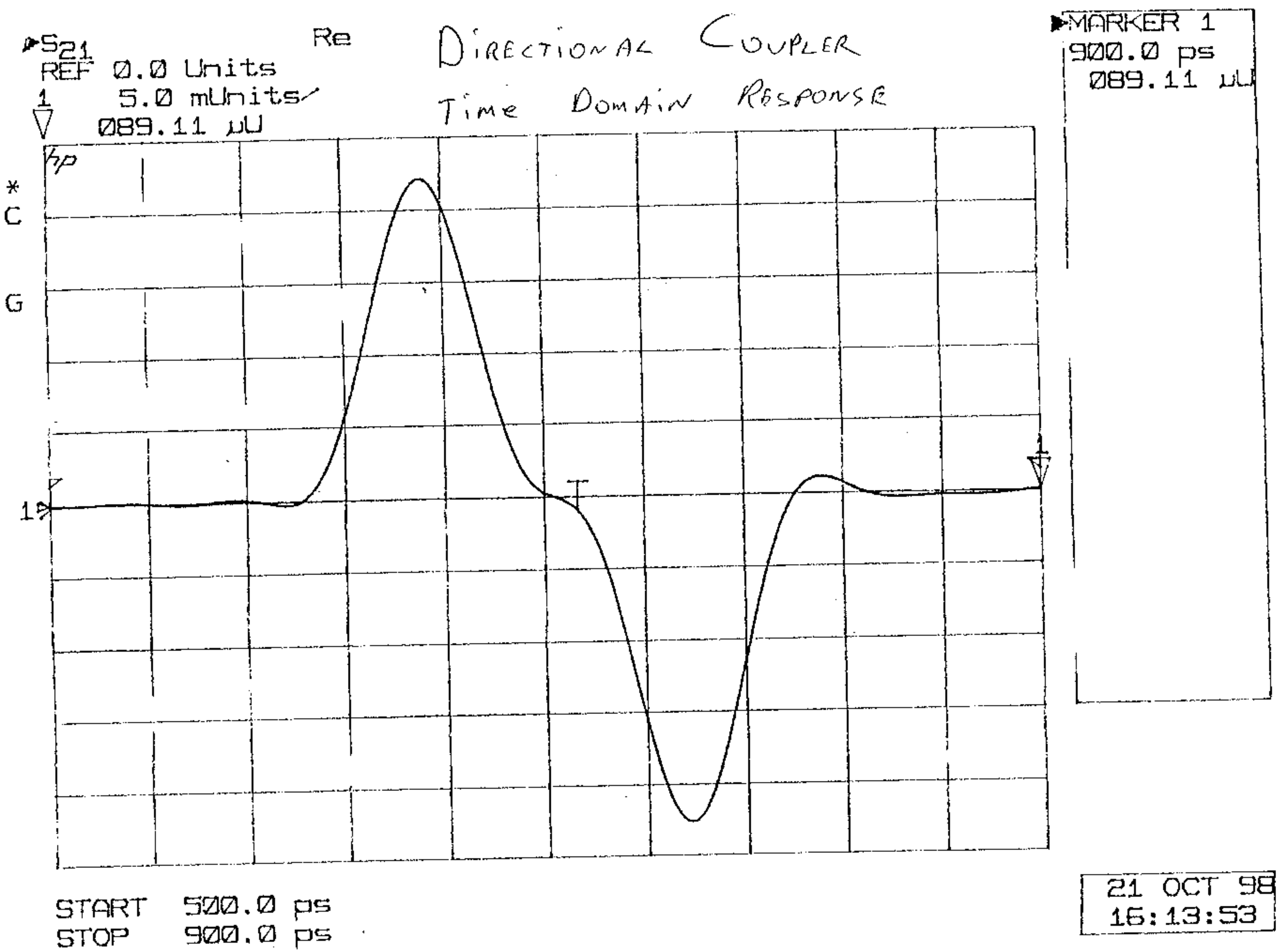
Ralph J. Pasquinelli
Fermilab

Directional Coupler

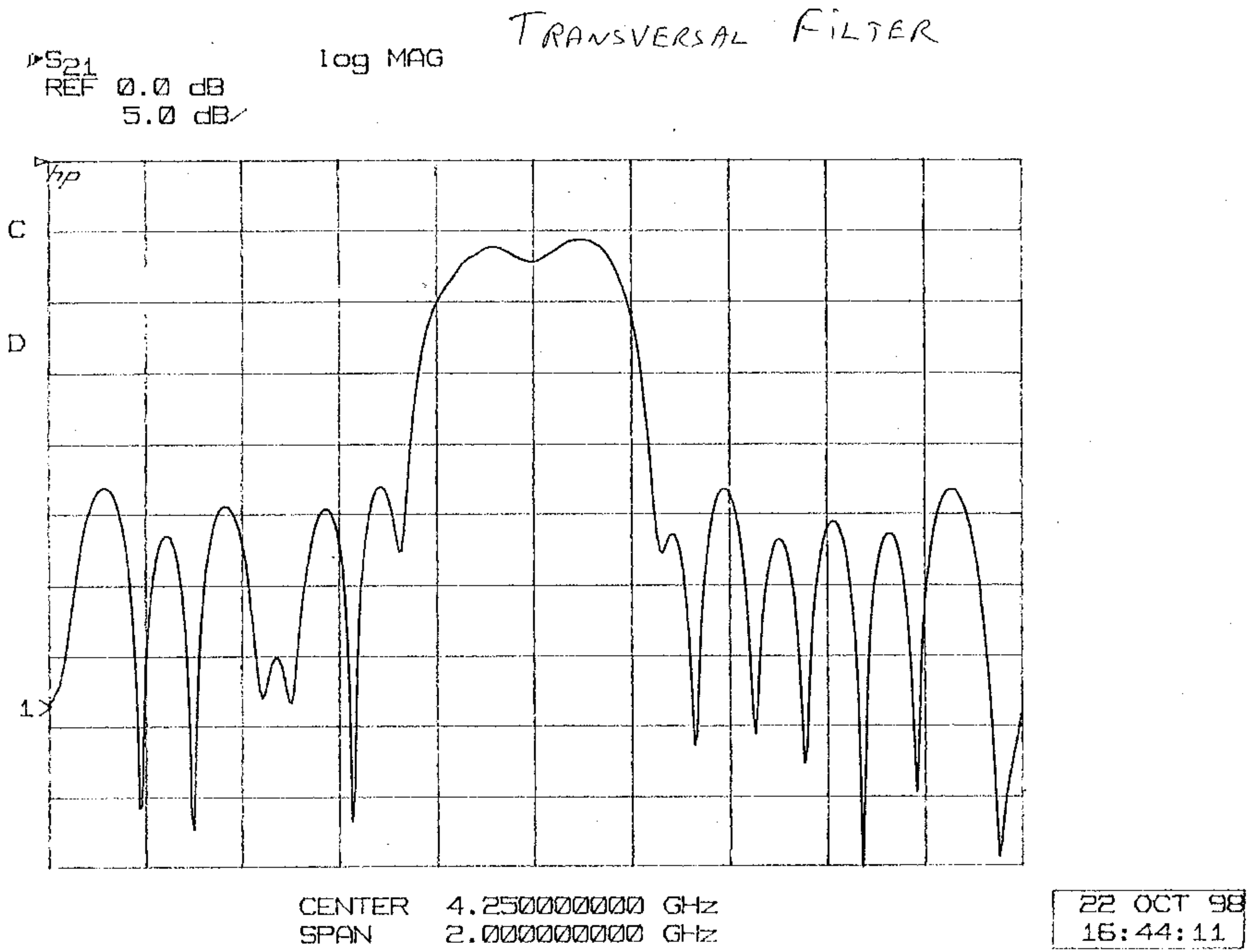


Ralph J. Pasquinelli
Fermilab


Directional Coupler
Frequency Domain Response
S21
REF -20.0 dB
5.0 dB/
log MAG
C
G
1
START 0.050000000 GHz
STOP 20.050000000 GHz
21 OCT 98
16:10:05

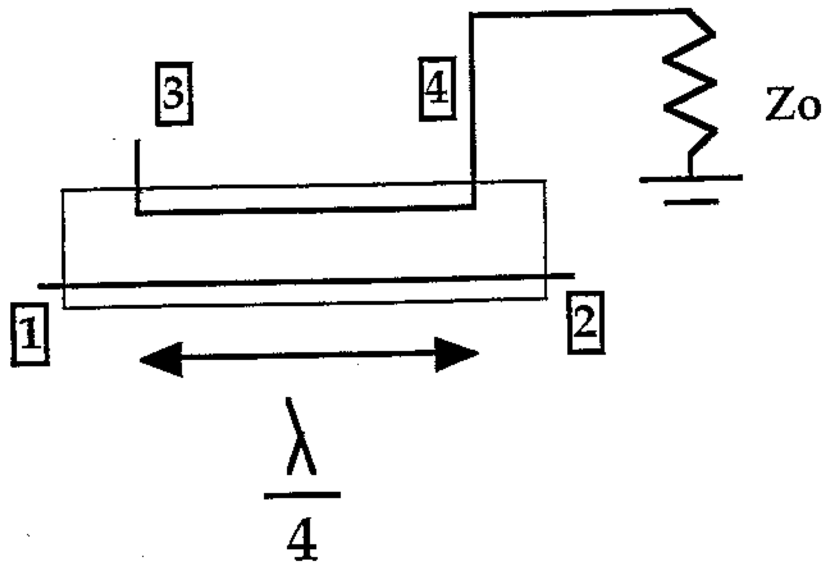


S21
Re
REF 0.0 Units
5.0 mUnits/
089.11 µU
DIRECTIONAL COUPLER
TIME DOMAIN RESPONSE
MARKER 1
900.0 ps
089.11 µU
hp
C
G
START 500.0 ps
STOP 900.0 ps
21 OCT 98
16:13:53



TRANSVERSAL FILTER
S21
log MAG
REF 0.0 dB
5.0 dB/
hp
C
D
1
CENTER 4.2500000000 GHz
SPAN 2.0000000000 GHz
22 OCT 98
16:44:11

Directional Coupler Characteristics



$$CouplingFactor = 10\log \frac{P1}{P3} dB$$

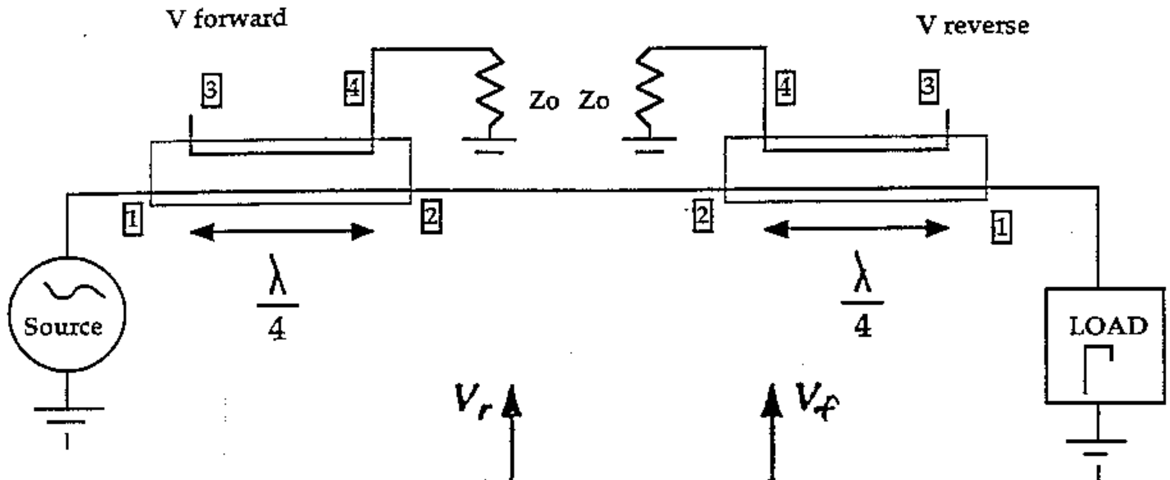
$$Directivity = 10\log \frac{P3}{P4} dB$$

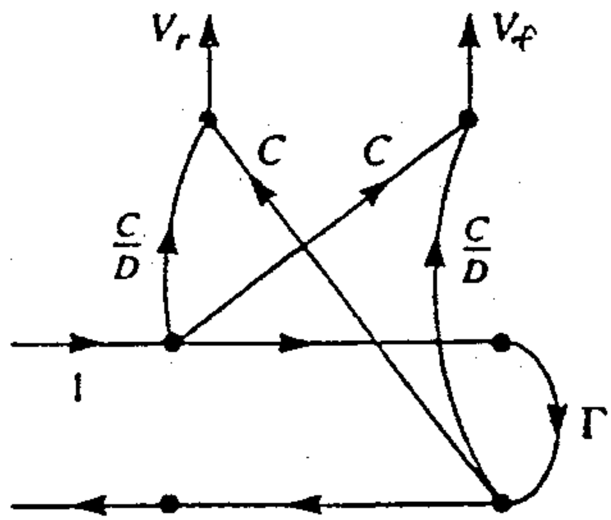
$$Isolation = 10\log \frac{P1}{P4} dB$$

Isolation = Directivity + Coupling factor

Ralph J. Pasquinelli
Fermilab

Reflectometer





$$V_f = C\left[1 + \frac{1}{D}\Gamma e^{j\theta}\right]$$

$$V_r = C\left[\Gamma e^{j\phi} + \frac{1}{D}\right]$$

$$\left|\frac{V_r}{V_f}\right|_{\substack{\min\\ \max}} = \frac{|\Gamma| \pm \frac{1}{D}}{1 \pm \frac{|\Gamma|}{D}}$$

Ralph J. Pasquinelli
Fermilab

90 Degree Hybrids

STRIPLINE CIRCUIT DESIGN

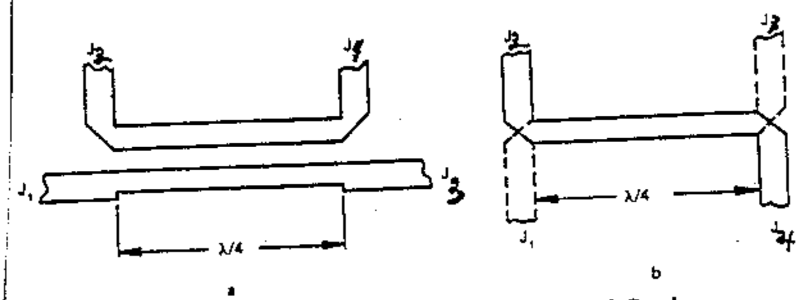
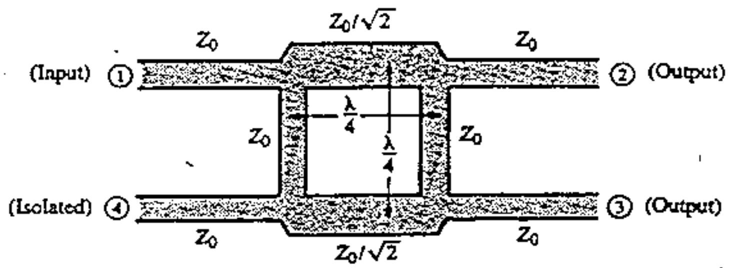


FIG. 5–1 Quarter-Wave Coupled-Line Directional Coupler Configurations



STRIPLINE CIRCUIT DESIGN

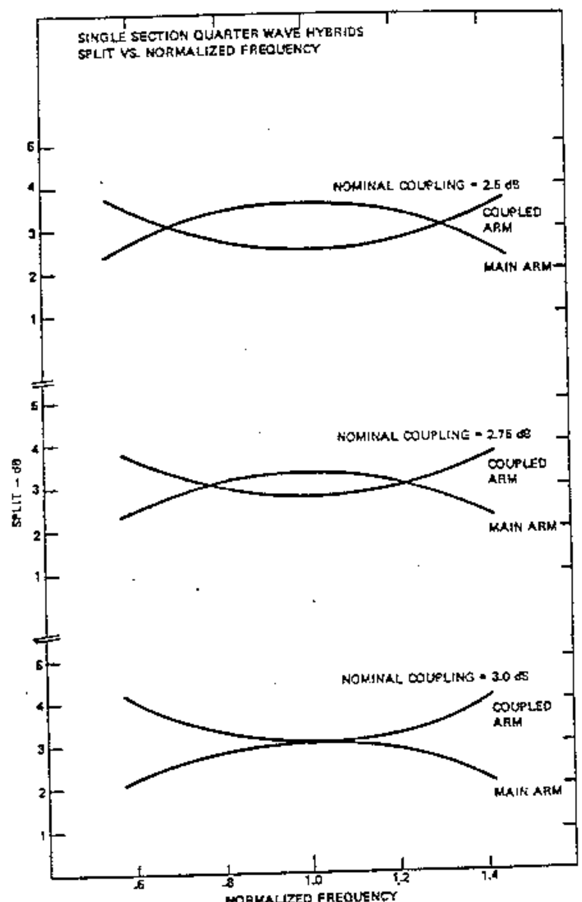


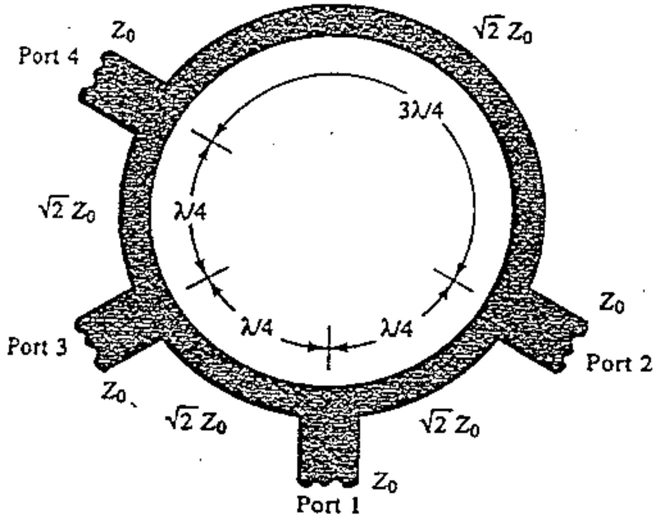
FIG. 5–5 Frequency Response for Several Single-Section Quarter-Wave 3.0 dB Hybrids for Varying Values of Nominal Coupling

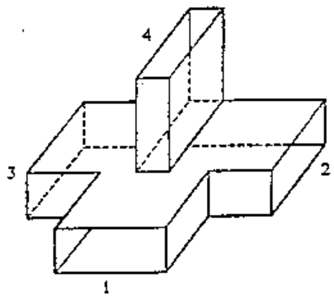
$$[S_{90}] = \frac{-1}{\sqrt{2}}\begin{bmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{bmatrix}$$

Ralph J. Pasquinelli
Fermilab

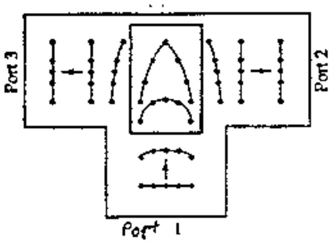
180 Degree Hybrids

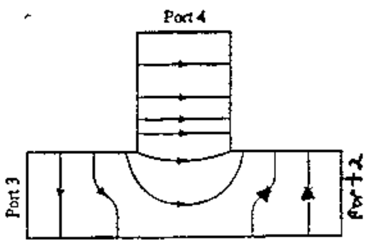
Ring Hybrid





Waveguide Magic Tee

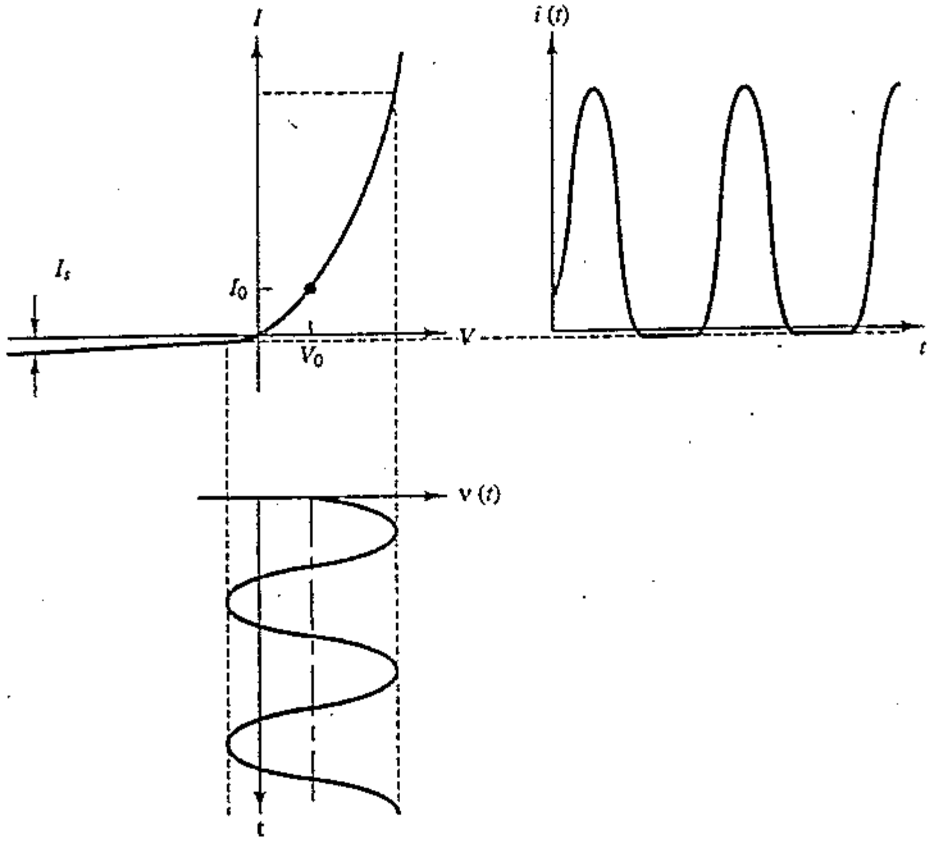




$$[S_{180}] = \frac{-j}{\sqrt{2}}\begin{bmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & -1 \\ 1 & 0 & 0 & 1 \\ 0 & -1 & 1 & 0 \end{bmatrix}$$

Ralph J. Pasquinelli
Fermilab

Diode Detectors



Square Law Detectors

$$i = a_o + a_1 v + a_2 v^2 + a_3 v^3 + \ldots$$

if

$$v = A\cos\omega t$$

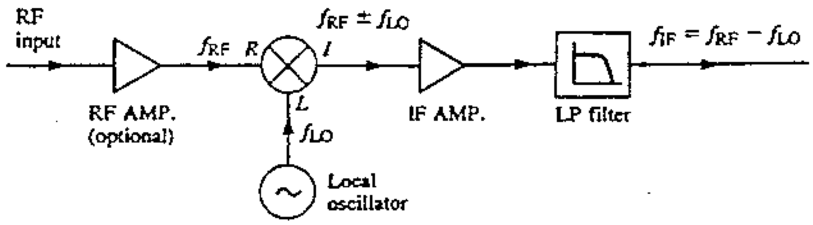
$$i = a_1(A\cos\omega t) + a_2(A\cos\omega t)^2$$

$$i = a_1(A\cos\omega t) + \frac{a_2 A^2}{2}(1+\cos 2\omega t)$$

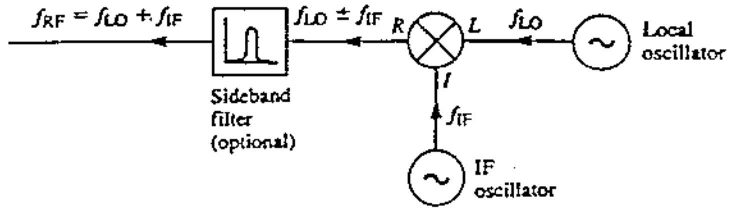
Ralph J. Pasquinelli
Fermilab

Mixers

$$V_{out} \approx k\, V_{signal} V_{LO}[\cos(\omega_{LO} - \omega_{signal})t - \cos(\omega_{LO} + \omega_{signal})t]$$

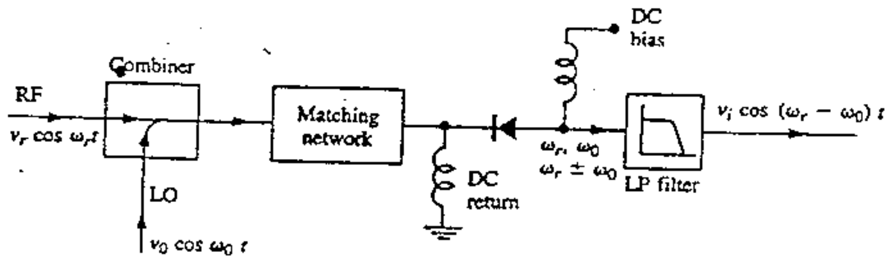


(a)



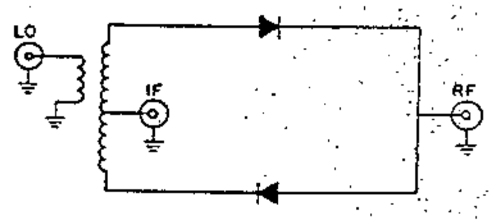
Single-ended

Simple design



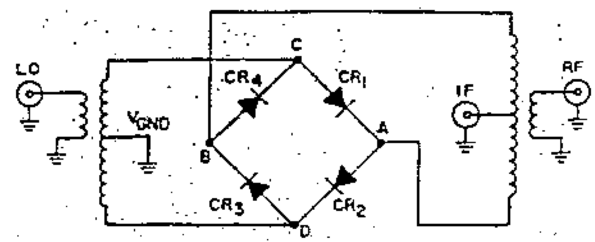
Single Balanced

Improved input match
Isolation between RF and LO



Double Balanced

Improved Isolation between all ports
Suppresses even harmonics of RF and LO
Low conversion loss



Ralph J. Pasquinelli
Fermilab

References

Microwave Theory and Applications, Stephen Adam , 1969

Handbook of Filter Synthesis, Anatol Zverev, 1967

Reference Data for Radio Engineers, ITT, 1974

Stripline Circuit Design, Harlan Howe Jr., 1974

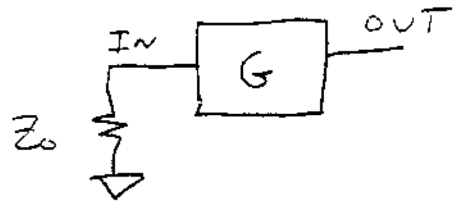
Microstrip Lines and Slotlines, Gupta, Garg, Bahl, 1979

Cablewaves Systems, Catalog 600, 1986

Ralph J. Pasquinelli
Fermilab

RJ Pagnanelli 1/22/99

NOISE FIGURE MEASUREMENTS



NOISE IN   NOISE ADD   NOISE OUT

$$\text{NOISE FIGURE} = NF = \frac{\dfrac{\text{SIGNAL IN}}{\text{NOISE IN}}}{\dfrac{\text{SIGNAL OUT}}{\text{NOISE OUT}}} = \frac{\text{SIGNAL IN}}{\text{SIGNAL OUT}} \times \frac{\text{NOISE OUT}}{\text{NOISE IN}}$$

FOR A LINEAR AMPLIFIER / DEVICE

$$\frac{\text{SIGNAL IN}}{\text{SIGNAL OUT}} = \frac{\text{SIGNAL IN}}{G \times \text{SIGNAL IN}}$$

$$\boxed{NF = \frac{\text{NOISE OUT}}{G \times \text{NOISE IN}}}$$

BY INSPECTION OF CIRCUIT

$$\text{NOISE OUT} = (\text{NOISE IN} \times G) + \text{NOISE ADD}$$

SOLVE FOR NOISE ADD

$$\text{NOISE ADD} = \text{NOISE OUT} - \text{NOISE IN} \times G$$

$$= NF(G \times \text{NOISE IN}) - \text{NOISE IN} \times G$$

$$= (NF - 1)\,\text{NOISE IN} \times G$$

$$\text{NOISE IN} = k T_o B$$

$$\boxed{\text{NOISE ADD} = (NF - 1) k T_o B G}$$

k = BOLTZMANN'S CONSTANT $= 1.374 \times 10^{-23} \dfrac{\text{joules}}{{}^\circ K}$

(joule = WATT SEC)

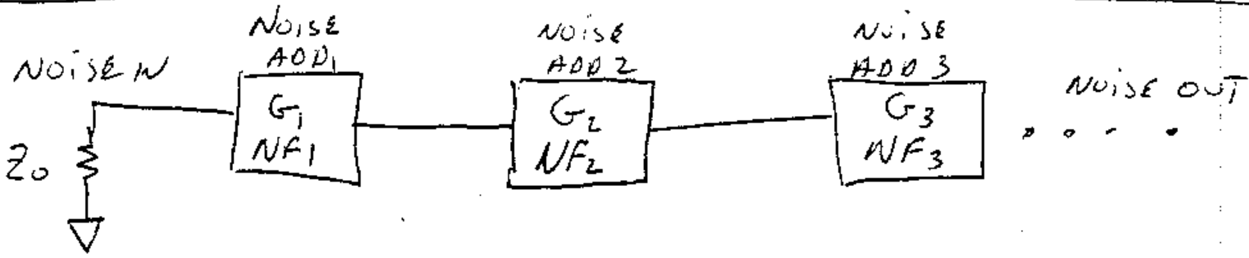
NOISE HAS UNITS OF WATTS

NOISE FIGURE MEASUREMENTS

PERFECT DEVICE ADDS ZERO NOISE

$$\text{NOISE ADD} = (NF-1)\,kT_oBG$$

$$NF = 1 \Rightarrow 10\log(1) = 0\,\text{dB}$$



NF_s = SYSTEM NOISE FIGURE

HERE GAIN SYSTEM $= G_1 \times G_2 \times G_3$

$$NF_s = \frac{\text{NOISE OUT}}{\text{NOISE IN} \times G_1G_2G_3}$$

$$NF_s = \frac{kT_oBG_1G_2G_3 + (NF_1-1)kT_oBG_1G_2G_3}{kT_oBG_1G_2G_3}$$

$$+ \frac{(NF_2-1)kT_oBG_2G_3}{kT_oBG_1G_2G_3}$$

$$+ \frac{(NF_3-1)kT_oBG_3}{kT_oBG_1G_2G_3}$$

$$\boxed{NF_s = NF_1 + \frac{NF_2-1}{G_1} + \frac{(NF_3-1)}{G_1G_2} + \cdots}$$

Signal to Noise and Dynamic Range Issues in System Design

Ralph J. Pasquinelli
Fermilab, Batavia, Illinois 60510*

Abstract

Study of signal to noise and dynamic range of systems is a very important part of engineering. The topic of signal to noise has been covered extensively in the literature, but not necessarily from a practical standpoint. Discussion of dynamic range issues is virtually missing from most fundamental texts. This paper will attempt to present practical ways of looking at system design. For completeness, the fundamental equations will be included, but the emphasis will be on real system implementation. The paper will draw extensively from actual system designs at Fermilab.

Sources of Noise

The random motion of electrons in materials due to temperature is a source of noise known as thermal noise. Electrons moving as a current through a solid state device or vacuum tube can generate shot or Schottky noise. In the case of an accelerator, shot noise can come from any source of accelerated charge, i.e. electrons, positrons, protons, and antiprotons. Sometimes the beam itself becomes the largest source of noise in an accelerator system. This noise can take the form of poor common mode rejection in a pickup device. All of these sources of noise tend to get in the way of analyzing the desired signal or waste precious kicker power in closed loop systems.

Figure 1 is a listing of the common equations used to calculate noise performance in systems. There is often confusion between the terms Noise Factor and Noise Figure. In this paper, Noise factor is the linear relation between noise on the output referred to the input and Noise Figure is the logarithmic relation expressed in dB.

Also included is the term of Effective Noise Temperature. Noise Figure and Noise Factor are commonly related to a room temperature of 290 deg Kelvin. A 3 dB Noise Figure or a Noise Factor of 2

*Operated by Universities Research Association under contract with the U.S. Department of Energy.

indicates double the input noise on the output, which for most terrestrial applications is the equivalent Effective Noise Temperature of 580 deg Kelvin. For space or cryogenic applications, the ambient temperature is not 290 deg Kelvin. The use of Effective Noise Temperature allows convenient arithmetic to calculate noise performance. For example if a pickup has an 80 deg Kelvin termination temperature and the preamp has an Effective Noise Temperature of 20 deg Kelvin, the front-end noise performance is 100 deg Kelvin.

$$\text{Noise Factor} = \frac{\text{Signal In}}{\text{Noise In}} \div \frac{\text{Signal Out}}{\text{Noise Out}} = \frac{\text{Signal In}}{\text{Signal Out}} \times \frac{\text{Noise Out}}{\text{Noise In}}$$

$$\text{Noise Power} = \text{K} \times \text{Temperature} \times \text{Bandwidth}$$

$$\text{Noise Factor} = \frac{\text{Noise Power Out}}{\text{K x Temperature x Bandwidth}}$$

$$\text{Noise Figure (dB)} = 10 \text{ x Log(Noise Factor)}$$

$$\text{Noise Figure (dB)} = 10 \text{ x Log}(\frac{\text{T2} - \text{T0}}{\text{T0}}) - 10\text{x log}(\frac{\text{N2}}{\text{N1}} - 1)$$

$$\frac{\text{N2}}{\text{N1}} = \text{Y\_Factor}$$

$$\text{Noise Figure (dB)} = 10 \text{ x Log}(\frac{\text{T2} - \text{T1xY}}{290\text{x(Y} - 1)} + 1)$$

$$\text{System Noise Factor} = \text{NF1} + \frac{\text{NF2} - 1}{\text{Gain1}} + \frac{\text{NF3} - 1}{\text{Gain1 x Gain2}} + \ldots\ldots$$

$$\text{Effective Noise Temperature} = 290 \times [\text{NF} - 1]$$

Figure 1. Basic noise equations. Note that Noise Factor (NF) is not in dB. K in all equations is Boltzman's constant = 1.38×10^{-23} watt sec/deg Kelvin. Temperature is in degrees Kelvin.

Noise Floor

The term noise floor is used often to describe the amount of noise power in a system. The noise floor of most systems depends on the thermal noise at the input. The noise energy is measured in joules and is Noise Energy = KT (joules or watt seconds)

where K is Boltzman's constant, 1.38×10^{-23} joules/deg K. For convenience of scaling, it is always desirable to note noise power in a 1 Hz bandwidth. If T is room temperature of 290° K, then

$$KT = 4 \times 10^{-21} \text{ joules} = 4 \times 10^{-18} \text{ milliwatt sec}$$

convert to dBm in one Hz bandwidth,

$$KTB = -174 \text{ dBm/Hz} \quad (\text{room temp of } 290^\circ \text{ K})$$

For an ideal spectrum analyzer at room temperature that has 1 MHz of resolution bandwidth, the noise floor would be 60-dB higher or -114 dBm. Inspection of figure 2 shows that an expensive spectrum analyzer has a noise floor of -91 dBm which is much worse than ideal. This of course is due to input losses, front-end mixers. etc.

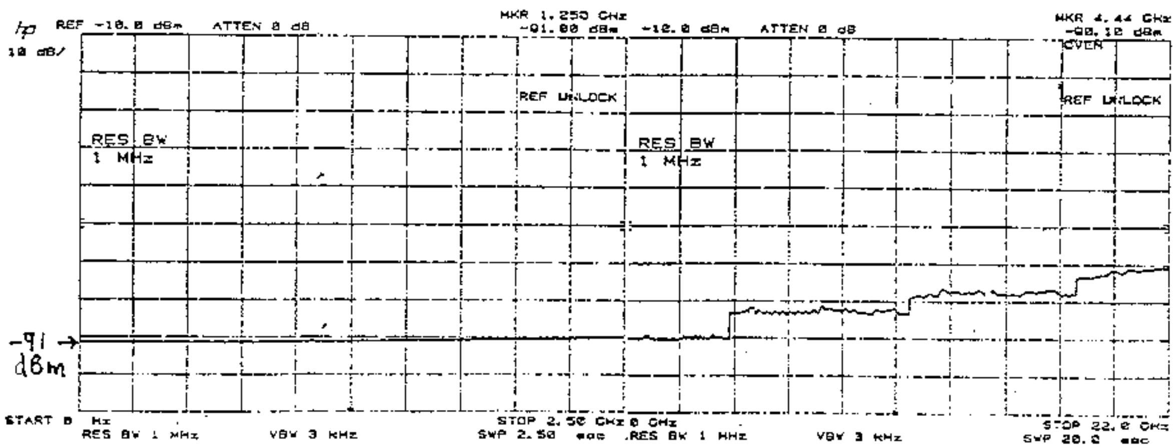


Figure2. Noise floor performance of HP 8566 spectrum analyzer. Span is from "DC" to 22 GHz, noise floor is -91dBm at best.

source are an excellent example to study Signal to Noise issues in system design. Figure 3 shows a simplified diagram of a closed loop system. In the case of stochastic cooling, the signal is the shot noise of the antiproton beam. The noise is the thermal noise floor of the system. (The beam also generates unwanted shot noise as well, but more about that later.) Because the source of antiprotons on our planet is non existent, we have to manufacture them by targeting protons on a copper target and collecting the 10 part per million yield. When this small beam current (nano-amps) goes through the pick up array, a small signal is developed on the order of 10 picowatts.

Let us take an example of a final system. Most of the cost in a feedback system is typically in the final high power amplifier. For this example, use the cost of $100 per watt. Careful analysis of the signal to noise ratio in a system can save money and provide the best performance. After all the gain calculations have been done, a system specification can be generated. Most machines also have a very limited amount of physical space to locate a kicker. All kickers have a maximum power capacity, so it is critical to make the best use of the installed power. As can be readily seen at this point, there are real restraints on building an actual system.

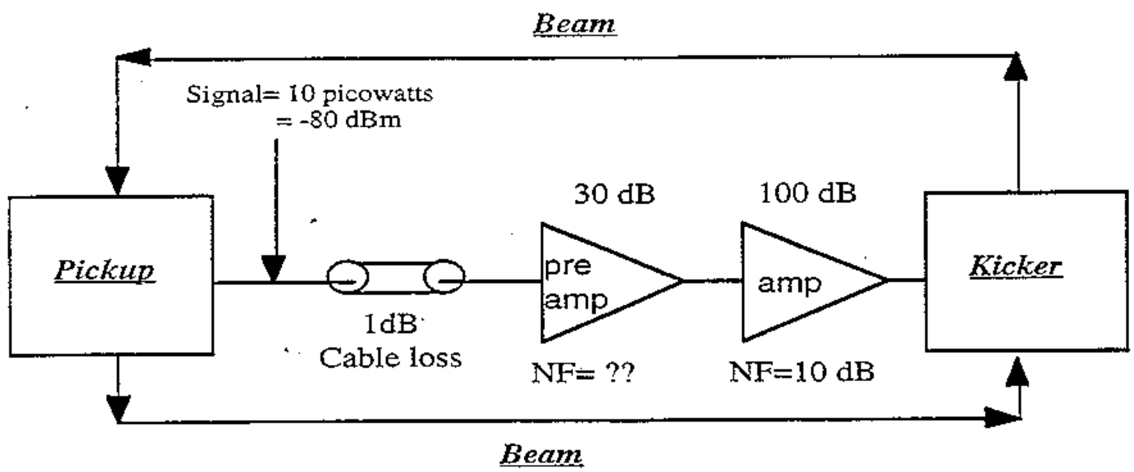


Figure 3. Typical stochastic cooling feedback system block diagram.

So here are the specs:

Amount of kicker signal power....................................100 watts, +50dBm

Amount of pickup signal power...............................10 picowatts, -80 dBm

System Bandwidth.. 1 GHz

Upon first inspection, a casual observer might just say, "That's easy, we just need 130 dB of gain and the project is over!" In reality, it is very important to carefully analyze the system and calculate the thermal noise power in the pickup with the given bandwidth. This is simply KTB where K is Boltzman's constant, T is the temperature in deg Kelvin, and B is the system bandwidth in Hertz. If we assume that our pickup will be at tunnel temperature of 27 deg C or 300 deg K, we can now find the thermal noise power. 4.14×10^{-12} watts or -83.8 dBm. We will analyze three cases of front-end hardware.

Case I will have a preamp that has a 3-dB noise figure. The cable at the input to the amplifier has 1 dB insertion loss and is the first stage in the gain chain. Using the equations presented earlier one can figure the noise factor of this system to be 2.52, noise figure of 4.01 dB, or effective noise temperature of 441 degrees Kelvin. Note here that the noise performance is dominated by the two first devices in the gain chain and that loss before the preamp directly degrades the noise performance. Multiplying the input by the 129-dB total gain will give a total of 79 watts of signal. Noise on the output referred to the input is the definition of noise factor, for 83 watts of thermal noise at the kicker and a signal to noise ratio (SNR) of -0.2 dB. Total power is 162 watts.

Case II involves going to the store and buying a more "state of the art" preamp that has a noise figure of 1 dB. The noise factor for this system is 1.58, noise figure of 2 dB, and noise temperature of 168 degrees Kelvin. At the kicker, we will see the same 79 watts of signal but now only 52 watts of noise for a SNR of 1.8 dB. Total power is 131 watts.

Case III will use the same 1 dB noise figure amplifier but put the pickup and preamp in a cryo environment of 80 deg K. We also decide to get rid of the 1-dB insertion loss cable because the preamp is now part of the pickup. The noise power on the input has a new KTB value of -89.6 dBm; the signal power remains unchanged. The noise factor of this system is 1.26, noise figure of 1.0002dB, and effective noise temperature of 75 deg Kelvin. In this system, the kicker will see 100 watts of signal and only 14 watts of noise for a SNR of 8.6 dB. Total power is 114 watts.

This performance comes at what cost. Remember that the system spec calls for 100 watts of signal, so for Case I & II an additional 26% of total power is necessary to go from 79 to 100 watts of desired signal. The cost comparison is based on front end and power costs. It is assumed that all the connections in between are the same in all cases.

Case I.	Preamp cost	\$500
	Power cost	\$20,500
	Subtotal	\$21,000
Case II	Preamp cost	\$1000
	Power cost	\$16,500
	Subtotal	\$17,500
Case III	Cryogenic cost	\$50,000
	Preamp cost	\$1000
	Power cost	\$11,400
	Subtotal	\$64,400

Note that in case III due to cryogenics, the cost is more than triple but cannot be ruled out. In the case of the stochastic cooling systems at Fermilab, the kicker power cost is between \$300 and \$500 per watt. For this reason, the cost of refrigeration saved dollars but more importantly provided the required performance.

Be aware also that the actual kicker may only have a power rating of 100 watts, or that there is limited space for adding extra kickers to handle the excess noise power, or there might be possible saturation of the kicker if ferrites are employed.

All amplifiers unfortunately reach saturation at some power level. In the case of the 100-watt amplifier, is 100 watts the linear output power, the one dB compression power or the saturated output power? As saturation is approached, the power amplifier also becomes a source of noise in the form of intermodulation distortion. For a wide band system, it is probable that odd order intermodulation products will be in the pass band as is shown in Figure 4. If the conditions of your system cannot tolerate the intermodulation noise it may be necessary to degrade your one hundred-watt amplifier to 50 watts or less making the dollars per watt increase accordingly.

Shot Noise Example

As an example of a noise source other than thermal, take for example an optical amplifier produced by BT&D. This device is an optical to optical amplifier; no electrical signal regeneration is required. The amplifier is similar to a semiconductor laser diode biased below threshold. The reflective coatings have been removed so that lasing does not occur. Amplification occurs (bi-directionally) when incoming photons create stimulated emission in the device junction. Figure 5 shows the shot noise performance of this device as a function of bias current: The larger the bias current the higher the gain, but at the expense of added shot noise. This device has been used to create a unity gain optical storage ring that functions as a recursive "brick wall" notch filter for bunched beam stochastic cooling in the Fermilab Tevatron.[2]

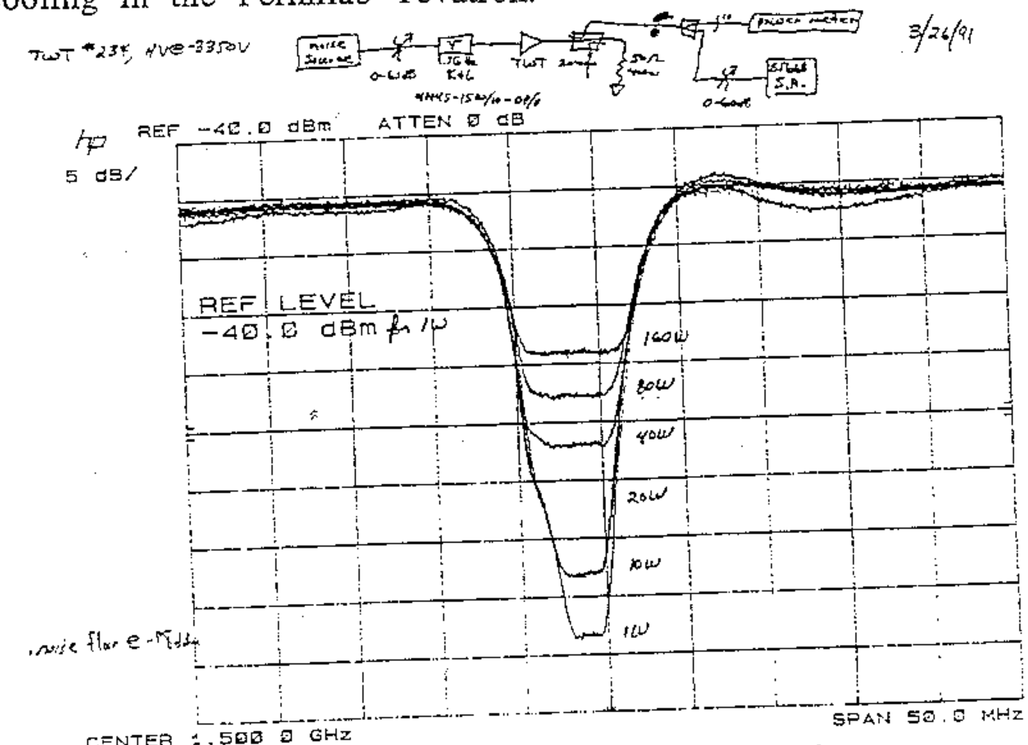


Figure 4. Odd order Intermodulation Distortion of a Traveling Wave Tube amplifier. Tube is driven with a notch filtered white noise source.

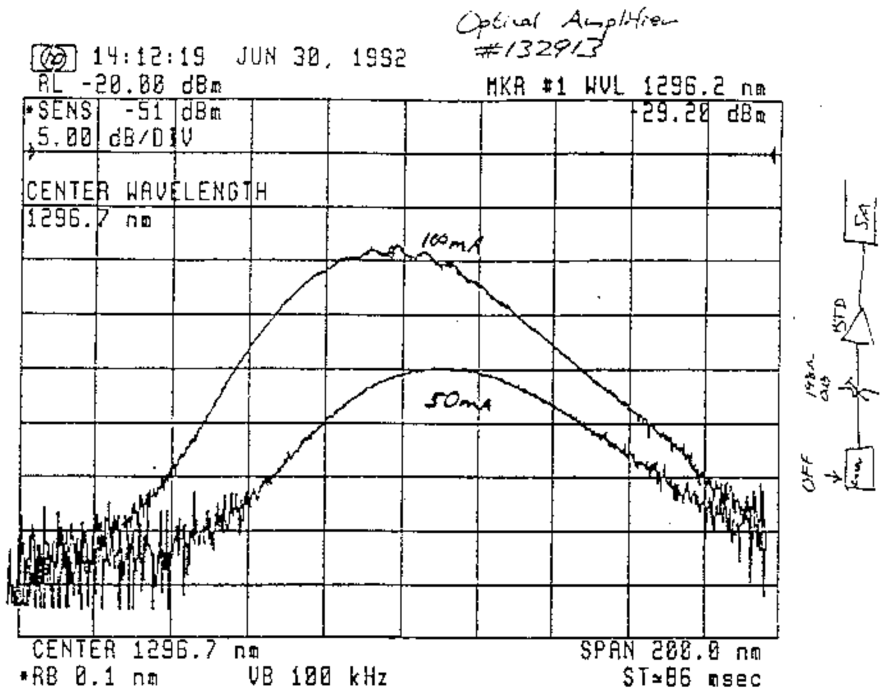


Figure 5. BT&D SOA3200 optical amplfier. Shot noise at the output is shown for two different bias points.

Beam Noise Example

In the accelerator environment, there are other sources of noise besides thermal and shot. The beam signal may itself be a huge source of noise. When attempting to analyze the transverse aspects of a beam signal, the longitudinal portion of the signal spectrum presents a "noise problem". The issue of dynamic range also comes up with this example. For a bunched beam circulating in an accelerator, the spectrum may very well look something like figure 6. The transverse information is in the betatron sidebands and is the useful signal. The large longitudinal line is unwanted noise that will also be amplified. Assuming flat gain in the pickup, a signal to noise ratio can be computed for this spectrum. Clearly, the unwanted longitudinal line dominates the power output. The longitudinal beam signal presents noisy watts that may or may not affect the beam in a feedback system or cause saturation of a preamplifier. Nonetheless, they are very expensive useless watts. How to get rid of them? There are two possible techniques used to reduce the longitudinal line, one is better common mode rejection in the pick up electrodes. This requires excellent mechanical tolerances and the best differencing circuit you can find or build. The second is to build some kind of filter that will only affect the longitudinal line leaving the gain and phase of the sidebands untouched.[2] Of course, complete elimination of the lines would require infinite common mode rejection, which is not a realistic expectation. Even with your best effort, there will always be some of the unwanted signal remaining. As shown with the thermal noise case, compromises are required to obtain the best possible performance.

Dynamic range

What is dynamic range? A careful search through the indexes of many engineering texts books comes up empty. Something so very important is never really taught at the basic level. In words, it is the ratio of the maximum available output power and the noise output power of a device at its rated gain.

Dynamic range = Maximum Power Out/Noise Power Out

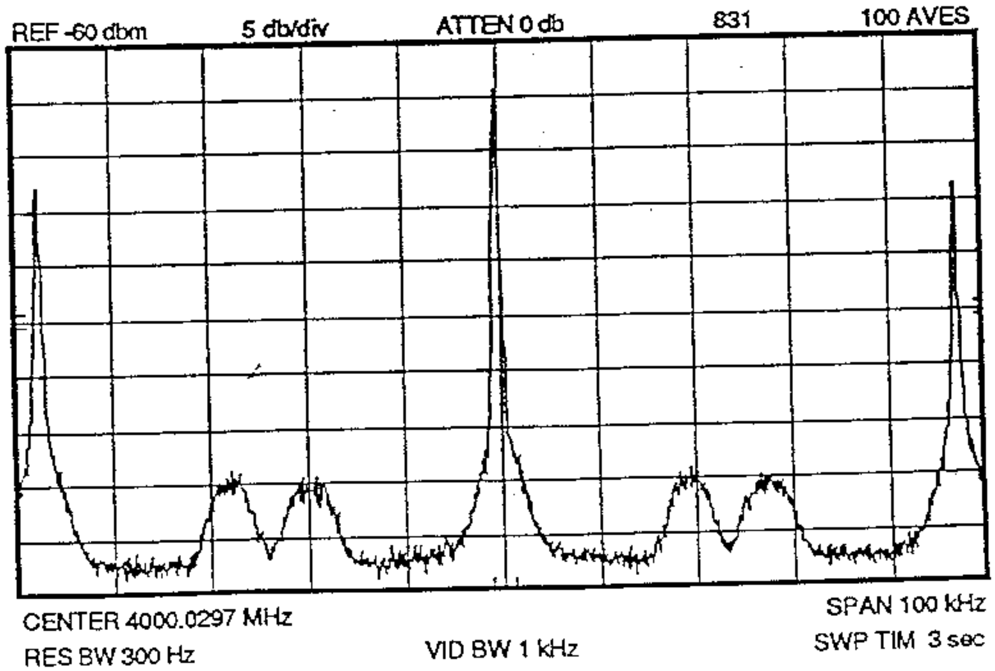


Figure 6. Spectrum of Bunched beam cooling signal in Fermilab Tevatron. The large coherent signal in the middle is unwanted longitudinal beam signal that behaves as system noise.

All devices are specified with a maximum input or output power over a specified bandwidth. Take for example an amplifier that is capable of delivering one watt of power over a wide bandwidth. At any single frequency, the unit should provide one watt. If the application is wide band, then only a fraction of a watt can be delivered at any given in band frequency. This in turn changes the dynamic range of the amplifier. A narrow band application will have a higher dynamic range over a wide band system using the same device. Figure 7 graphically shows the effect.

A practical example of this effect can be seen in a very wide band optical transmitter such as the Ortel 5515B. This unit has 12 GHz of modulation bandwidth but a maximum drive power level of 10 milliwatts. As can be seen in Figure 8, the noise floor also increases as modulation bandwidth increases (KTB again). The dynamic range can be maximized with narrow band modulation, but will suffer as Bandwidth increases.

Peak vs. Average Power

The tools available to investigate signals are network and spectrum analyzers for the frequency domain and oscilloscopes for the time domain. Each domain has its pluses and minuses. Peak amplitude will cause devices to saturate and is easily observed in the time domain. For example figure 9 is a sequence of three oscillographs of the same signal viewed in three different places in the gain chain of the bunched beam cooling system of the Tevatron[3]. There are two adjacent bunches in the accelerator, the main bunch and an unwanted satellite. The unwanted satellite does give us a good indication of amplifier saturation. It is evident that saturation has started after the second amplifier. This same information in the frequency domain is not obvious as a spectrum analyzer displays the average power. Similarly, when the information is frequency related such as Schottky signals in figure 6, the frequency domain is preferred.

The signals used in the previous example were actually the same signal. Using both domains prove to be valuable tools in understanding the system.

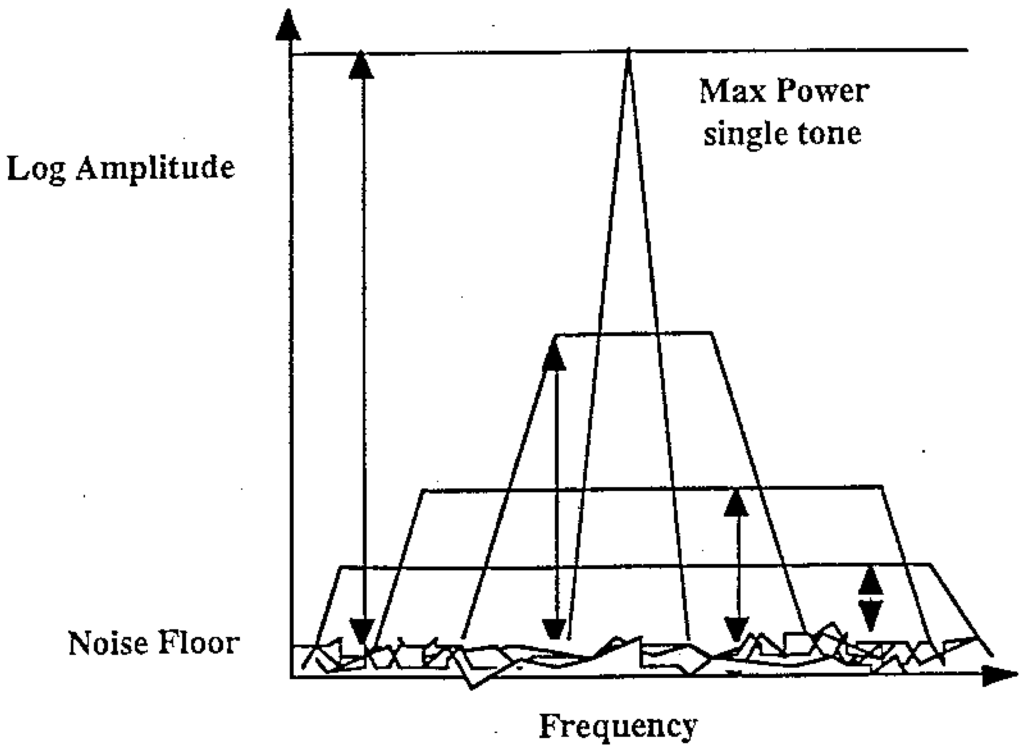


Figure 7. Narrow band operation has larger dynamic range than wide band for the same device due to maximum power limitations.

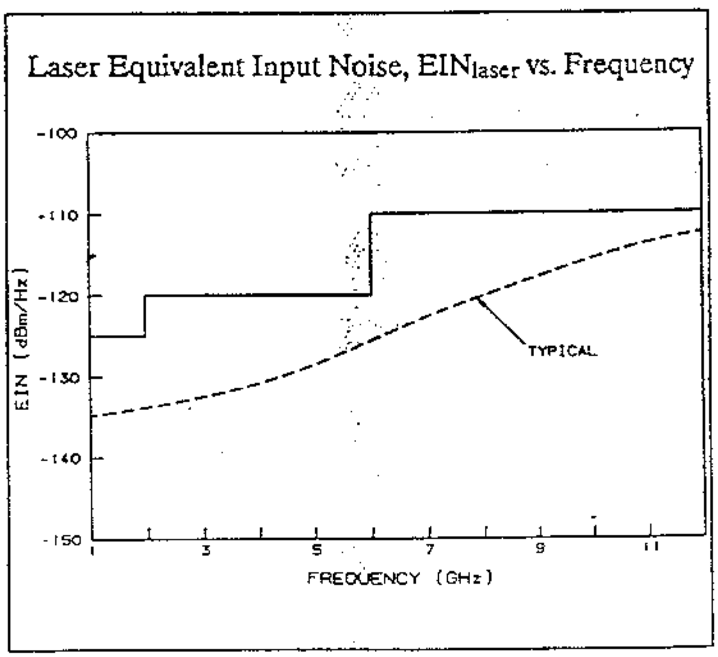


Figure 8. Noise performance of Ortel 5515B laser transmitter.

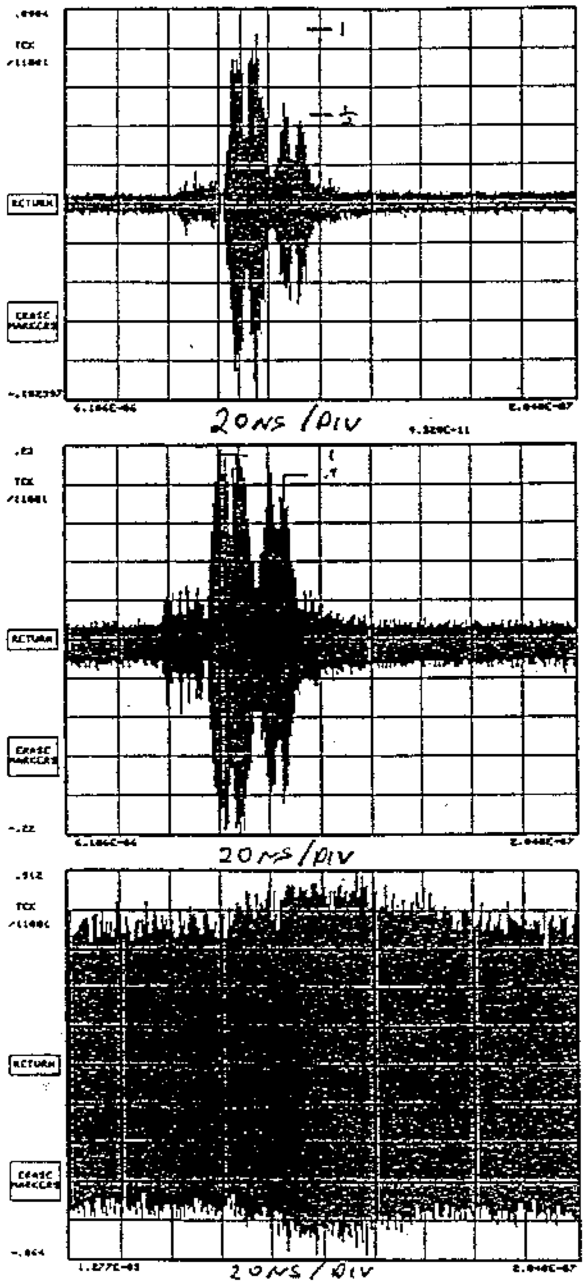


Figure 9. Oscilloscope traces of Bunched Beam Cooling system in the Tevatron. Top trace is signal after first preamp. Middle trace same signal after second amplifier. Bottom trace same signal after final TWT amplifier.

Gating

In the case where the signal is not continuous (most accelerators have beam with a bunched structure) gating can be used to improve the signal to noise ratio. Figure 10 shows the Schottky signal of the bunched beam cooling system in the Fermilab Tevatron where the beam is in six bunches. The time between bunches contributes only to the noise of the system. Gating around the bunches dramatically improves the signal to noise ratio by a factor of the gating duty cycle.

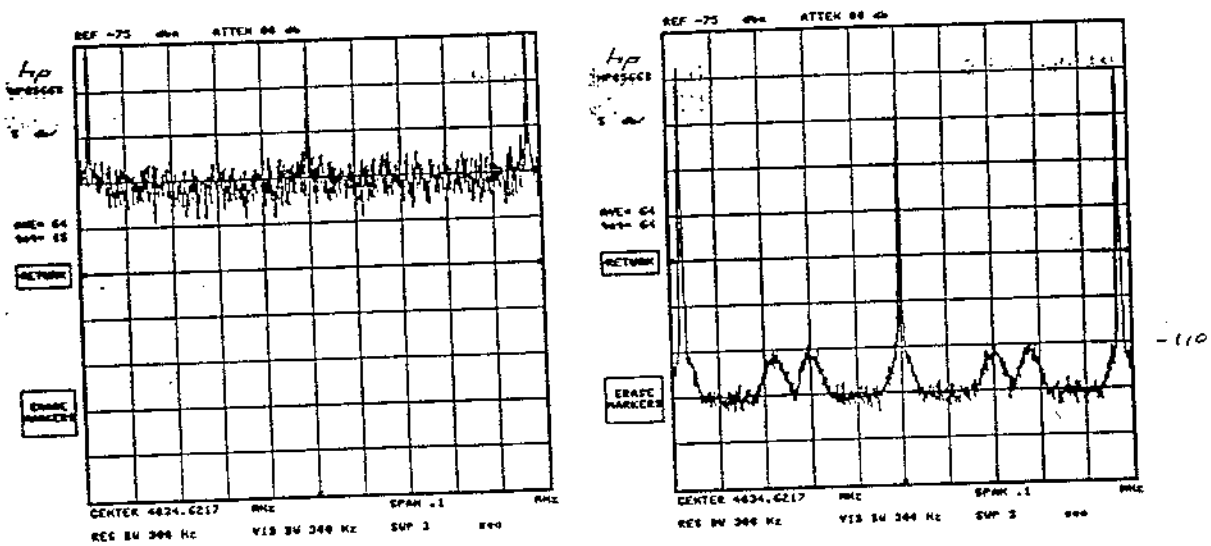


Figure 10. Effect of gating on signal to noise ratio.
Left ungated Bunched Beam Cooling signal.
Right same signal with gating.

Digital Connection

A few words about signal to noise and dynamic range when working with digital circuits. The world of logic is predominantly base two. A factor of two is 6 dB in the voltage that an analog to digital converter (ADC) sees. Hence, an ADC of ten-bit resolution has a maximum dynamic range of 60 dB. If the analog input to the device has a SNR of 48 dB, the remaining two bits (12 dB) of the converter are meaningless. Matching the digitizing number of bits to the signal being digitized is important. Those bits are very expensive to create and manipulate. Don't buy more than you need. It's very popular today to boast about the number of bits a system can digitize, but that is not the significant side of the ADC.

Conclusions

Much of what has been presented is common sense. Careful inspection of system requirements and good engineering practice at the beginning of a project can result in better performance at a lower cost.

References

1. J. Marriner, Review of the Physics, Technology and Practice of Stochastic Beam Cooling, IEEE Particle Accelerator Conf. 1987, Vol. 3, pp 1383-1387.

2. R. J. Pasquinelli, Electro-Optical Technology Applied to Accelerator Beam Measurement and Control, IEEE Particle Accelerator Conf. 1993.

3. G. Jackson et al., A Test of Bunched Beam Stochastic Cooling the Fermilab Tevatron Collider, Proc. of 1991 IEEE Particle Accelerator Conf., pp. 1758-1760.

4. J. Hardy, High Frequency Circuit Design, Reston Publishing Co. 1979.

5. S. F. Adams, Microwave Theory and Applications, Prentice Hall , 1969.

6. BT&D corporation SOA 3200 data sheet, Wilmington, DE 19803.

7. Ortel corporation 5515B laser transmitter data sheet, Alhambre, CA. 91803.